MICROWAVE TECHNIQUES

NAVSHIPS 900,028

Prepared by

NDRC—Division 14
RADIATION LABORATORY
Massachusetts Institute of Technology

Published by

BUREAU OF SHIPS NAVY DEPARTMENT

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D. H. CLARK,
Rear Admiral, USN,
Chief, Bureau of Ships.

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TABLE OF CONTENTS

Item	Page
Letter of Transmittal	III
Acknowledgments	IV
Table of Contents	V
I. Introduction	1
II. Transmission Lines	1
1. Coaxial Lines	1
a. Stub supports	1
b. Bead-supported coaxial transmission lines	4
c. Connectors for stub-supported lines	5
d. Sizes of lines	7
e. Rotary joints in coaxial lines	9
f. Flexible cables	12
2. Coaxial Lines and Wave Guides	17
a. Propagation in wave guides	17
b. Modes in general	18
c. Modes in rectangular guides	19
d. Modes in circular guides	22
e. Modes in coaxial lines	22
f. Coaxial lines versus wave guides	24
g. Resonant sections	25
3. What Can Be Done With Wave Guides	26
a. Obstacles in wave guides	26
b. Bends and twists in wave guide	29
c. Joints in wave guide	30
d. Plungers	32
e. T-joints in rectangular guide	32
f. Effects of dirt, water condensation, and silver plating	37
g. Effect of dielectric	38
h. Wave guide transformers	39
i. Coaxial line to wave guide transformations	44
j. Rotary joints	47
III. R-F Components of 10 cm. and 3 cm. Radar Systems	48
1. Magnetron Coupling	48
2. The TR Box	50
a. Purpose and theory of the TR box	50
b. 10 cm. TR boxes	53
c. 3 cm. TR's	58
3. Local Oscillators and Mixers	67
a. Local oscillators	67
b. Mixers	71
c. Crystals	76
4. Antenna Matching	83
5. R-F Heads	84
6. Pressurizing	86
7. Spinner Housing Design Considerations	88

Item	Page
IV. Test Equipment and Measurements	91
1. Slotted Sections	91
a. Standing wave detection in general	91
b. Coaxial slotted sections	92
c. Wave guide slotted sections	94
2. Probes, Probe Detectors, and Indicators	96
a. Crystal probes	97
b. Bolometer probes	102
c. Thermocouple probes	106
d. Meters and amplifiers	106
3. Variable Impedance Matching Transformers	112
a. Coaxial line tuners	112
b. Wave guide tuners	116
c. R-f power divider	120
4. Wavemeters and Echo Boxes	121
a. Wavemeters	121
b. Echo boxes	128
5. Test Bench R-F Oscillators	131
a. General purpose generators	131
b. Modulated systems	133
c. Power supplies	134
6. Power Measurements	135
a. High-level power measurements	135
b. Low-level power measurements	145
7. Terminations	148
a. Coaxial matching terminations	148
b. Wave guide matching terminations	150
8. Attenuators	153
a. Cut-off attenuators	153
b. Dissipative attenuators	155
9. Test Equipment, Larger Units	159
a. R-f spectrum analyzer	159
b. R-f envelope indicator	162
c. Test sets	165
10. Measurements	167
a. *Q* measurements	167
b. F-m bench measurements	170
c. Loss measurements	173
d. Impedance measurements	176
V. Appendix	181
1. Power Lost in Reflected Wave	181
2. Guide Wavelength vs. Free Space Wavelength	182
3. Circular Impedance Chart	183
4. Coaxial Line Data	184
5. Rectangular Wave Guide Data	184
6. Round Wave Guide Data	185
Index	187

I. INTRODUCTION

The original purpose of this book was to relieve the members of Group 53 of the burden of explaining apparatus to newcomers and other members of the Radiation Laboratory who were not familiar with the work of Group 53. The equipment and methods discussed here are developments of many groups in the Radiation Laboratory as well as outside organizations.

None of the material here presented is new in the sense that it cannot be found elsewhere, because all of the subjects have been exhaustively treated in various publications. What is new is that all of these subjects are covered in one volume and are treated from the point of view that they are tools for further work rather than ends in themselves. Too many topics are treated to cover each one in complete detail. For that reason, only the more important facts about each subject are presented.

A knowledge of conventional line theory and impedance charts is essential to the understanding of the radio-frequency components of a radar system. That the reader has at least a speaking acquaintance with standing wave ratios, quarter- and half-wave length transformers, shorted sections of line, etc., has been assumed in this handbook. However, a short discussion of these subjects is given in section IV, 10d.

II. TRANSMISSION LINES

1. COAXIAL LINES

a. Stub Supports.

The lowest and most used mode of oscillations in a coaxial line is that in which the electric vector is radial from the center conductor to the outer conductor. (See figure 1.) Coaxial lines carrying this mode are the only ones discussed in this text. In some places reference will be made to the fact that more difficulty is encountered in achieving the desired results with large-sized lines. This is because other modes can be propagated in a coaxial line if the diameter exceeds a certain limit.

The largest line which can be used without excessive higher mode troubles is one in which the average of the circumferences of the center and outer conductors is less than a wavelength. With lines larger than this, a mode which has a variation of electric field in the angular coordinate can be propagated as well as the normal mode. In this case it is very difficult to know exactly what is going on.

FIGURE 1.—Electric field in normal mode.

The center conductor of a coaxial line must be supported so that it remains coaxial with the outer conductor at all times. To do this at longer wavelengths, one normally uses coaxial lines filled with a dielectric medium. The electrical losses in the best dielectrics known make it prohibitive to use these for long runs of line carrying microwaves. An alternative method of support is by the use of dielectric beads as discussed in the next section.

Mechanically and electrically the best coaxial lines for microwaves are the so-called stub-supported lines which are used to a great extent at these wavelengths. A stub consists of a short-circuited section of coaxial line tied across the normal transmission line as in figure 2.

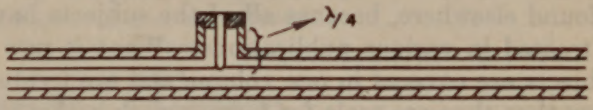


FIGURE 2.—Quarter wavelength stub support.

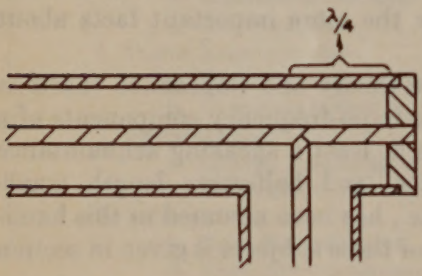


FIGURE 3.—Stub support for right-angle bend.

This stub is in parallel with the line and, if it is made exactly the equivalent of a quarter-wavelength long, it will give no reflection at that frequency. Such a stub acts as an insulator because it has an infinite reactance when it is a quarter-wave long. It may also be used to make a right-angle bend as in figure 3.

It is possible to make direct right-angle bends in coaxial lines, but in lines which are fairly large, considerable mismatch results unless special constructional features are used. One quite successful method to get around this mismatch is shown in figure 4.

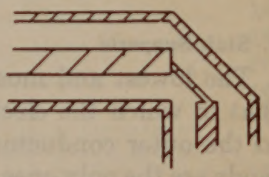


FIGURE 4.—Direct right angle bend.

Since a quarter-wave stub may be thought of as a loaded resonator, it will be perfectly reflectionless at only one wavelength. At longer wavelengths, the stub is less than a quarter-wave and therefore shunts the line with an inductive susceptance, and at shorter wavelengths, the stub is more than a quarter-wave and so shunts the line with a capacitive susceptance. To compensate for this

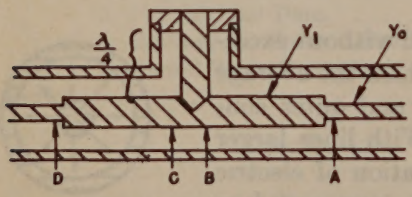


FIGURE 5.—Broad-banded stub support.

effect, it is possible to place two stubs in the line an odd number of quarter-waves apart so that their respective reflections tend to cancel.

Another way to compensate for the effect of a single stub is to combine it with a half-wave transformer consisting of an enlarged section of center conductor symmetrically placed about the stub, as shown in figure 5.

The rectangular admittance charts of figure 6 illustrate the action of this device. These charts are simply rectangular coordinate paper on which the ordinate gives positive or capacitive susceptance in the upward direction and negative or inductive susceptance in the downward direction from center. The abscissa represents conductance increasing from 0 to the right. Y_0 represents the characteristic admittance of the normal line and Y_1 the characteristic admittance of the section of line forming the half-wave transformer. If the line

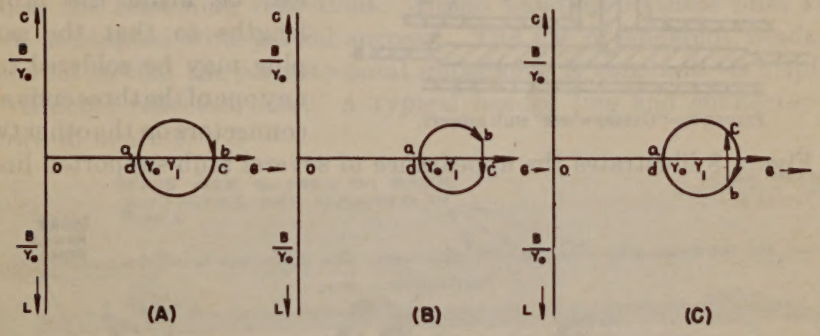


FIGURE 6.—Explanation of the broad-banded stub support by means of rectangular impedance charts.

is terminated in Y_0 at a, at the design frequency, where the transformer is exactly one-half wavelength and the stub the equivalent of a quarter-wavelength long, going to the left through the transformer simply takes us around the circle *abcd* in figure 6 (A). Thus the line to the left of the transformer is matched.

At longer wavelengths we follow figure 6 (B). From Y_0 at a, we go less than a quarter-wave around Y_1 to b. Then, the stub adds an inductive susceptance taking us to c. Another less than a quarter-wave section of line of surge admittance Y_1 returns us to Y_0 at d.

At shorter wave lengths than in figure 6 (B), we use figure 6 (C). Here a to b is more than a quarter-wave long, taking us below the real axis. Then the stub adds capacitive susceptance taking us to the point marked c, and finally c to d, being again longer than a quarter-wave, returns us to Y_0.

This kind of stub can be made very broad having less than 1 percent mismatch over $\pm$ 15 percent wavelength band. It is quite simple and so has found considerable application in microwave transmission lines.

Right angle stub supports as in figure 3 are not reflectionless at any wave length when the diameter of the line becomes large. It is necessary in these cases to add some kind of matching transformer to get reflectionless transmission. It is possible to considerably increase the band width by combining the transformer and stub in such a manner that their frequency sensitivities tend to cancel. It is also possible to combine such a matching transformer with the band-broadening transformer of figure 5.

Figure 7 shows such a combination in which Y_2 is greater than Y_1. This stub is one which can be used as either a right angle bend or as a straight through device depending upon whether the shorting plug is soldered into *c* or *a* or *b*. The stub can be made the proper lengths so that the same plug may be soldered into any one of the three ends and connectors on the other two.

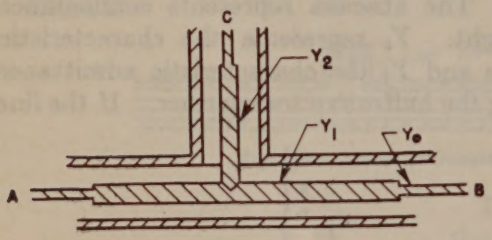


FIGURE 7.—"Gozanywhere" stub support.

Figure 8 illustrates the appearance of several stub-supported lines.

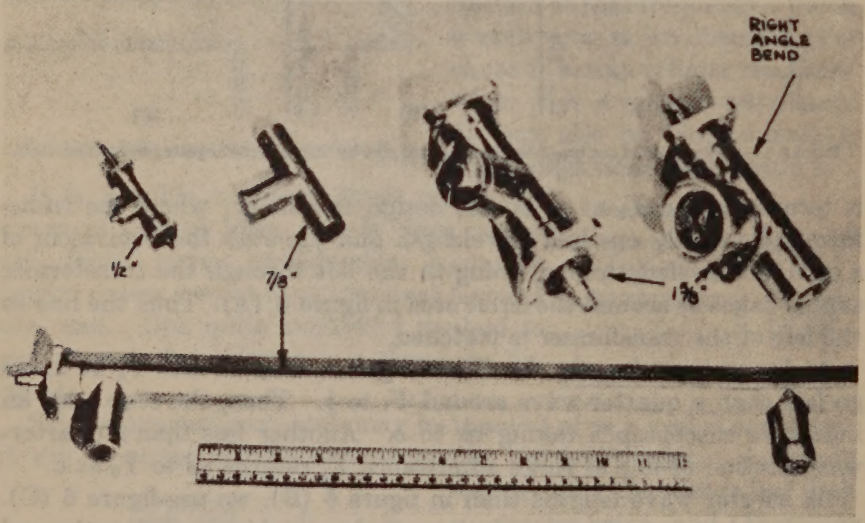


FIGURE 8.—Stub supported lines.

b. Bead-Supported Coaxial Transmission Lines.

The first rigid transmission lines for use in the 10 cm. band were constructed by using plastic (polystyrene etc.) beads to support the inner conductor. In the lower frequency ranges ceramic and plastic bead-supported lines have been widely used. Since these lines suffer from low break-down voltage (due to the air-dielectric interface), poor

mechanical strength and not too good electrical properties (in the microwave region), they are now almost entirely replaced by stub-supported coaxial lines and by wave guides. In the lower frequency regions the tendency is very strongly in favor of larger coaxial solid dielectric cables to replace the bead-supported coaxials.

However, since historically the first systems at Radiation Laboratory were made with "R. L. Standard" $\frac{5}{8}$" bead-supported line, they are worthy of brief mention. A set of associated r–f components was developed for use with this line. The lines are easily fabricated in the laboratory, and since annealed copper tubing is used, they can easily be bent to any desired shape.

The problem of reflection of beads is an important one at microwave frequencies, and many attempts were made to overcome it. In the "Lawson line" the beads are so arranged in groups that reflections tend to cancel, and the groups in turn arranged in sets of groups to cancel the remaining reflections. Broad banding of these lines has been attempted with partial success. The use of undercut beads—undercut so that the point-to-point impedance is constant—is helpful in reducing reflection loss. A typical beaded line and connector is shown in figure 9.

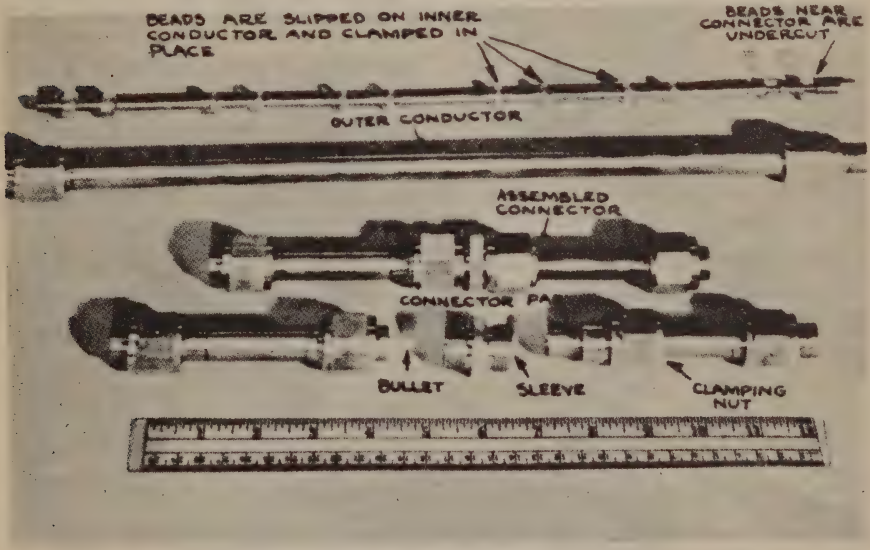


FIGURE 9.—$\frac{5}{8}$-inch beaded line and connector.

c. Connectors for Stub-Supported Lines.

In order that there be no reflection from a union between two similar coaxial lines, the connector must offer no discontinuity to the line. Both the inner diameter of the outer conductor and the outer diameter of the inner conductor must remain constant as a wave travels from one section of line to another.

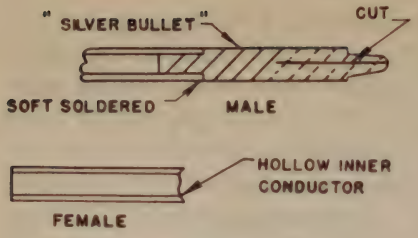


FIGURE 10.—Inner conductor connector.

The inner conductors of ⅞-inch stub-supported lines are connected by means of a silver-plated "bullet" (so-called because of its shape) which is soldered into the male half of the connector and which fits inside the hollow inner conductor on the female half. (See figure 10.) The front half of the bullet has two lengthwise cuts which form equal sectors that can be sprung apart, as shown in figure 11. *If the bullet fits loosely in the female half, the four prongs should be sprung apart so that a good contact is made.*

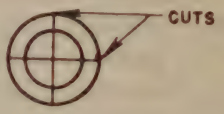


FIGURE 11.—Front view of "bullet."

The two outer conductors (fig. 12) meet in a taper: The male taper is slightly more pointed than the female, so that when they are drawn together, a very good contact is made at the tip of the taper. Outside the taper is a rubber gasket which makes the connection pressure tight. The two halves are drawn together and held firmly by the two flanges which are screwed together. The tapped holes are in the flange on the female half only.

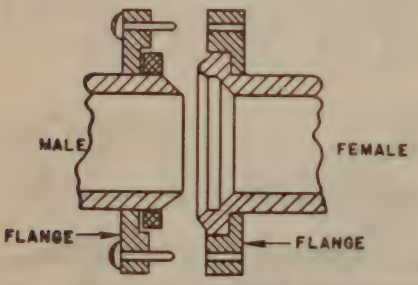


FIGURE 12.—Outer conductor connection.

The complete connection is made as follows: The coaxial line is cut off squarely and the parts shown in figure 13 assembled as indicated in figure 14. Exactly 1 inch is added to the line by the connector pieces.

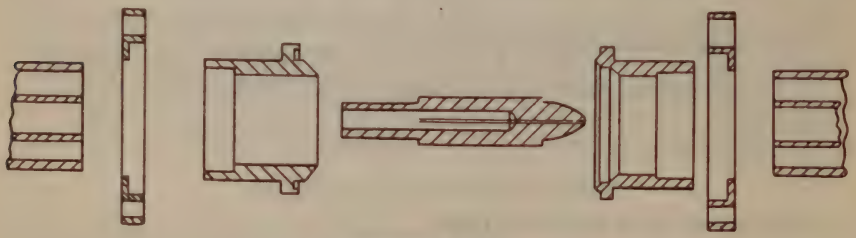


FIGURE 13.—The separate parts of a connector.

The dimensions of the connector are chosen so as to leave a 1/64-inch crack where the bullet meets the inner conductor of the female half (point *A* in figure 14). If the maximum error in manufacture is

±1/64-inch, there will be at *A* either a perfect fit or a 1/32-inch crack. The latter will produce (theoretically) a standing wave ratio of 1.03. If the bullet is made too long, the inner conductor will buckle when the connector is drawn together.

All of the stub-supported lines have connectors similar to the one described above. The 1⅝-inch connector has hollow bullets, but this and other differences are minor and unimportant.

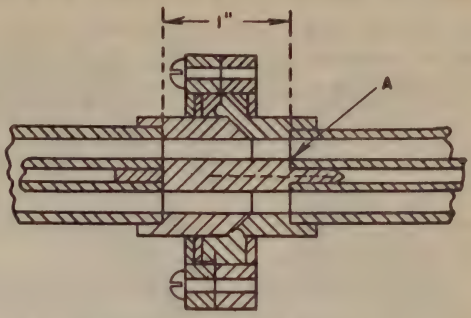


FIGURE 14.—Complete connector.

d. Sizes of Lines.

There have been at least four sizes of r-f lines in general use. All of these sizes have a nominal impedance of 50 ohms. The size most widely used is the ⅞-inch stub-supported line with ⅜-inch inner conductor. (See figure 15.) A line of this size has a theoretical power-carrying ability of 1.12 megawatts, assuming breakdown at

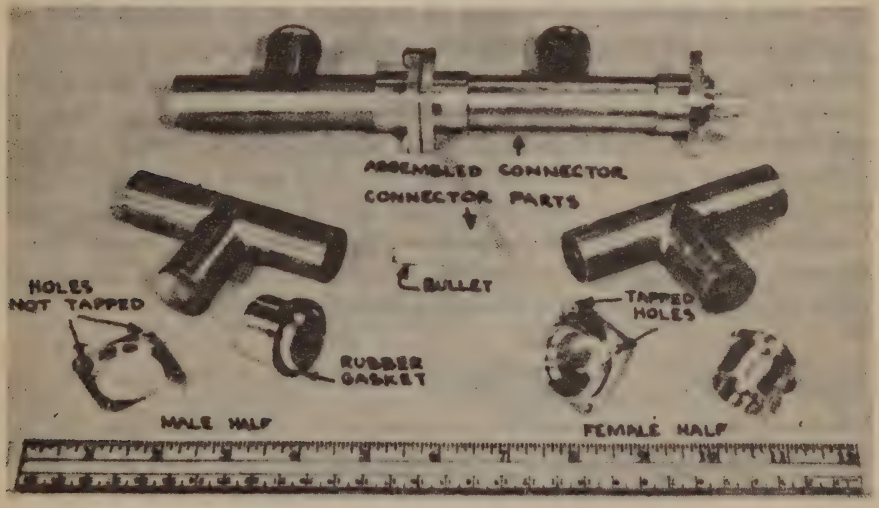


FIGURE 15.—Connector for ⅞ inch stub-supported lines.

30,000 volts per centimeter. The presence of stubs reduces this value. Powers of 600 kilowatts have been carried by this line but it is not recommended for service use for powers in excess of 200 kilowatts.

A line having an outside conductor ⅝-inch in diameter with a ¼-inch inner conductor has been used. This line has been bead supported. It has been shown theoretically that such a line would

not be good for powers in excess of 73 kw., again assuming 30,000 volts/cm. field. The smaller size reduces the ⅝-inch line power carrying ability by a factor of 0.44 as compared with an ideal line of ⅞-inch size. An additional factor of approximately 0.15 arises from increased fields in the small air spaces which are almost certain to occur between bead supports and the center conductor. Lines in service have given trouble and burned out at powers of about 50 kw.

A line having an outer conductor 1⅝-inch in diameter and an inner conductor ⅝-inch in diameter is now being used with megawatt powers. This is the largest permissible size of coaxial line for use in the 10 cm. region.

A $\frac{5}{16}$-inch line has been used for low-power applications. A ½-inch stub-supported line has been used in the 3 cm. region, and is the largest diameter suited for use at this wavelength.

Transforming from one size of line to another is usually accomplished by the use of either a tapered section or quarter wavelength transformers. For example, in transforming from a line A to a line B, where A and B represent different sizes, a short connecting section can be used whose dimensions at one end match the dimensions of A and at the other end match those of B, with a gradual taper in between. Although all the lines in use have a nominal impedance of 50 ohms, they have individual differences of as much as 9 ohms, so that an impedance transformation is also necessary. Hansen has shown that the match with a tapered section can be best accomplished if the length of taper is an integral number of half wavelengths. A connector can also be a quarter wavelength of line whose characteristic impedance is the geometric mean of those of the two lines being connected. This arrangement, however, can handle less power and is more frequently sensitive than a connector made up of two quarter wavelength transformers. The quarter wavelength sections of this latter connector must be such as to satisfy the following relations:

$$\frac{Z_0'''}{Z_0'} = \left(\frac{Z_0'}{Z_1}\right)^2 = \left(\frac{Z_T}{Z_0'''}\right)^2$$

where Z_1 is the characteristic impedance of the input line, Z_T the terminating impedance, Z_0' the characteristic impedance of the quarter wavelength section next to Z_1, and Z_0''' that of the section next to Z_T. (See figure 16.)

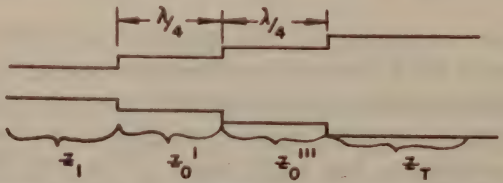


FIGURE 16.—Transformer made up of two quarter wavelength sections.

Tapered sections and double quarter wavelength transformers are illustrated in figure 17. Both types are satisfactory over a 20-percent band.

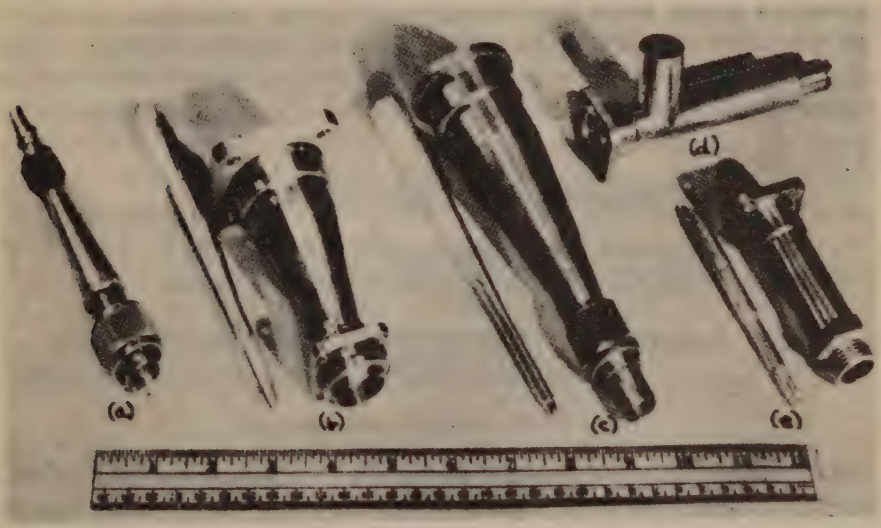


FIGURE 17.—Connectors between different sizes of coaxial lines. (a), (b), and (c) are tapered sections: (d) and (e) are double quarter wavelength transformers.

e. Rotary Joints in Coaxial Lines.

(*1*) *Contact type.*—This type is simple in theory but ranges from difficult to impossible in practice. For satisfactory performance the contacts must be good and the geometry of the joint must be such as to preserve the geometry of the line impedance. Usually this is done by slotting one of the contact surfaces on both inner and outer conductor so as to give a springy "finger" contact. (See figure 18.)

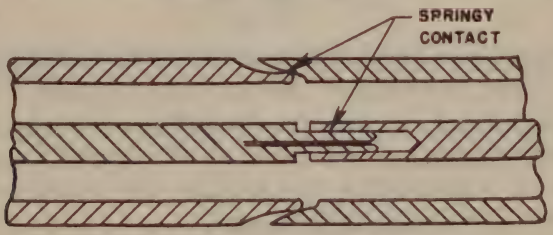


FIGURE 18.—Contact coupled coaxial rotary joint.

While these contacts may perform satisfactorily at low speeds for a time if carefully made, they are difficult to make right and deteriorate due to wear and (unless protected) corrosion.

(*2*) *Capacity type.*—If two similar coaxial lines are placed end to end, there will be enough capacity between the conductors of the two lines to couple some energy through the gap. One of the lines could then rotate with respect to the other. The capacity between the lines

may be increased by increasing the area of the adjacent surfaces, as shown in figure 19.

One would suppose that the longer the length l and the smaller the clearance d, the lower the capacitive reactance and hence the more efficient the coupling. However, there are limits on both. The clearance d is, of course, limited by mechanical considerations since actual contact must be avoided as discussed above. Increasing l does

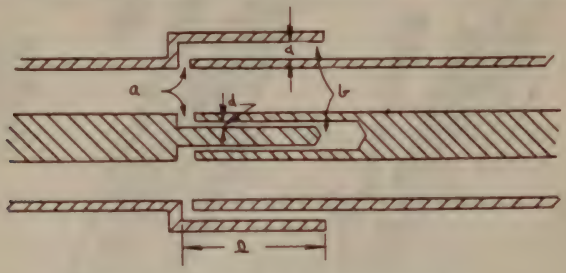


FIGURE 19.—Capacity coupled coaxial rotary joint.

improve matters, until a length of one-quarter wavelength is reached. Then the capacitive effects of additional elements become out of phase with the first-quarter wavelength. This is easily understood from line theory as follows: At the points b the impedance is very large. On the center rod there will be a very slight end capacitance, while on the outside of the line there is a large radiation resistance. Both are essentially open-circuit terminations. For such an open-circuit terminated line we know that the impedance at the points a will decrease to zero as l approaches $\lambda/4$, then increase to infinity again as l approaches $\lambda/2$. The expression is

$$Z_a = -jZ_0' \cot \frac{2\pi l}{\lambda}$$

Here Z_0' is the characteristic impedance of the coupling sections of line and, of course, decreases as the clearance decreases. This formula shows that if $l=\lambda/4$, Z_a is zero regardless of the clearance d. However, because of manufacturing tolerances on the length l and because of broad bands of wavelengths, the smaller the clearance (and hence Z_0') the slower Z_a will depart from zero.

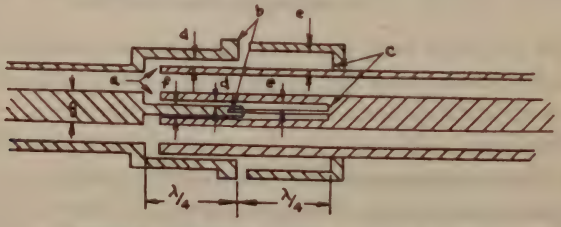


FIGURE 20.—Improved capacity coupled rotary joint.

An improved version of this joint is sketched in figure 20.

Additional lengths *bc* have been added to both outer and inner couplings. In the outer line the object is to suppress leakage of power by putting a high impedance (choke *bc*) in series with the radiation resistance at *b*. This also allows us to place a housing with bearings, pressure seal, etc., around the entire rotating joint without disturbing the open-circuit termination of *ab*. In the case of the inner conductor, the object of the section *bc* is to furnish a means of aligning the center conductors and maintaining the clearance *d*. The contact occurs at *b*, and is in series with the high impedance choke *bc*, which limits the current to practically zero. Hence no pains need be taken to make good contact here.

It will be noted that the gaps *e* are larger than the gaps *d*. Just as gap *d* should be as small as practical (see above), gap *e* should be as large as practical. This gives the desired highest attainable impedance at *b*, as can be seen by the equation:

$$Z_b = +jZ_0'' \tan \frac{2\pi l''}{\lambda}$$

where Z_0'' and l'' are the characteristic impedance and length of the choke *bc*.

As to the actual form of the capacity coupling elements, one finds several other versions, especially on the outer coupling (see figure 21). In each case *ab* and *bc* are each electrically equivalent to a quarter wavelength. Type (A) is being used on most of our present joints.

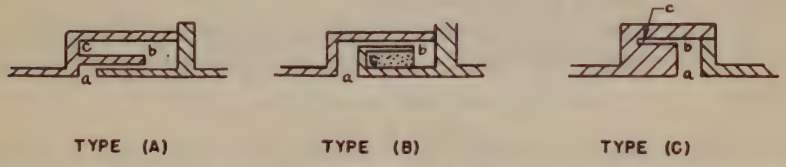


FIGURE 21.—Different forms of outer conductor coupling elements. Type (A)—*bc* folded back over *ab* Type (B)—*bc* filled with dielectric and folded back under *ab*, *ab* partly radial. Type (C)—*ab* radial.

Type (B) was used on earlier ⅝-inch joints, but the presence of dielectric was undesirable. Type (C) has been used in some 3 cm. band joints.

One would naturally worry about spark-over in these chokes. It turns out, however, that as the gaps are decreased, the voltages decrease about proportionally, so that the maximum gradient remains essentially the same as in the main line. This is not quite true, however, as regards fields at the center rod *f*. The field here is greater than the field at the center rod *g* of the main line by exactly the factor *g*/*f*. This is independent of clearance *d* except inasmuch as increasing *d* forces *f* to become smaller and hence makes matters worse.

Since this is the highest field in such a rotating joint, one may say that its power-handling capabilities are below that of the main line by a factor $(f/g)^2$.

As an example, take one of our standard $\frac{7}{8}''$ line rotary joints. Here $g=\frac{3}{8}''$ and f is about $\frac{1}{4}''$. Hence $(f/g)^2=(\frac{2}{3})^2=\frac{4}{9}\approx\frac{1}{2}$ is the power reduction factor.

The frequency sensitivity of this joint may be judged as follows: for $f=\frac{1}{4}''$, $d=\frac{1}{32}''$ Z_0' (inner) is 13 ohms. Assuming for the moment that the impedance at b is infinite

$$Z_a \text{ (inner)} = -j\ 13 \cot \frac{2\pi l}{\lambda} \text{ ohms}$$

if we increase λ by 10 percent, this becomes

$$Z_a = -j\ 13 \cot 81.9^\circ \sim -j\ 1.9 \text{ ohms}$$

The finite characteristic impedance of the choke section bc increases this figure by a factor of about

$$1+\frac{Z_{0_{(ab)}}}{Z_{0_{(bc)}}} \sim 1+\frac{13}{28} \sim 1.5$$

This gives $Z_a=3$ ohms, and is capacitative (as shown by $-j$). 10 percent decrease in wavelength would give about 3 ohms inductive reactance. Because the outer conductor chokes have the same clearances but larger diameters, they are only about one-third as frequency sensitive as the inner conductor chokes. Hence 10 percent off midband they give only about 1 ohm reactance.

In the sketch as drawn, these two reactances would add to give Z_a (outer) $+Z_a$ (inner)$\cong$4 ohms. The resulting standing wave ratio would be

$$r \approx \frac{50+4}{50} \approx 1.08$$

If convenient in design, one could make this still better by spacing the gap in the center conductor a quarter wavelength along the line from that in the outer conductor. This would give a reactance of $3-1=2$ ohms, and a standing wave ratio $r \approx 1.04$.

Figure 22 is a photograph of two rotating joints for stub-supported $\frac{7}{8}$-inch coaxial line.

f. Flexible Cables.

Flexible cables for transmitting radio-frequency power can be divided into two types: beaded cables and solid dielectric cables. Beaded cables are made by slipping "fish spine" beads (usually polystyrene) over the inner conductor, and then braiding fine copper wire over the beads to form the outer conductor; solid dielectric cables are made by extruding a flexible dielectric over the inner

conductor, again with a braid to form the outer conductor. The inner conductor may be either a single conductor or a composite of stranded wires. Beaded cables are easy to make but are not adaptable for quantity production. Also they have a low electrical breakdown strength and change characteristics with use. For these reasons we will limit this discussion to the much more available and useful solid dielectric cables.

Construction.—The inner conductor is usually made of copper wire or copper weld wire. In some large-size cables copper tubing is used. The braid is fabricated from copper wire which is occasionally tinned or plated. The insulating material for the low-loss cables made by General Electric and Amphenol is a mixture of polyisobutylene and cyclized rubber in the ratio of $^{65}/_{35}$. Cables made by Federal Telephone

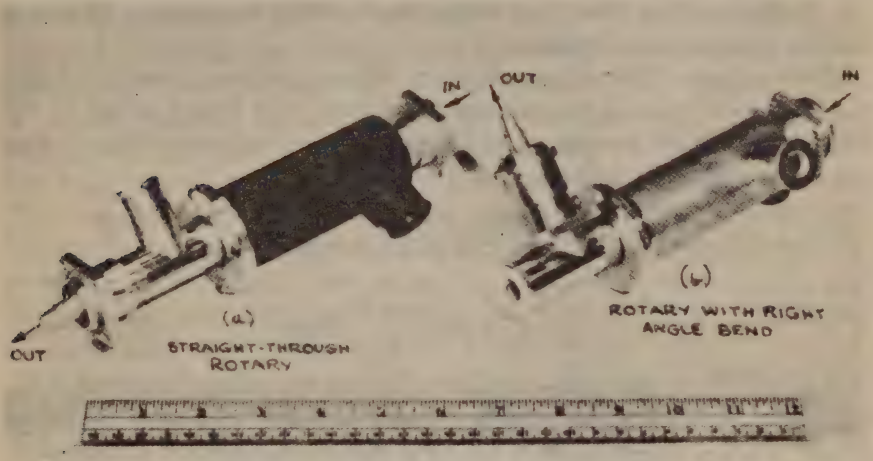


FIGURE 22.—Rotary joints for ⅞ inch stub supported line.

& Radio Corporation are a mixture of polyisobutylene and polystyrene with 2 or 3 percent of cyclized rubber. British cables use an insulation of polyethylene (sometimes compounded with a small percentage of polyisobutylene for flexibility). Polyethylene is now manufactured in this country and much wider use of it in cables is anticipated.

These materials all have a dielectric constant of nearly 2.25. Their dielectric power factor, which is a measure of the electrical loss introduced by their use, is about 0.0012 for the General Electric and Amphenol material, about 0.0008 for the Federal material and close to 0.0004 for polyethylene.

The outer jacket is generally made of vinyl chloride plastic, *Vinylite*, *Flamenol*, etc. The average cable is safely flexible in the temperature range of −20° C., to +70° C., and this range is being extended by the use of improved materials. The average cable is fairly rugged

and can stand continued abuse without failure. The Navy often requires the addition of a steel braid "armor" over the vinyl jacket.

Impedance.—The surge or characteristic impedance of solid dielectric coaxial cables is given quite accurately by the theoretical expression:

$$Z_0 = \frac{138}{\sqrt{k'}} \log_{10} \frac{b}{a} \text{ ohms}$$

where k' is the dielectric constant and a and b are the inner and outer conductor diameters. If air is trapped in the cable while the dielectric is being extruded the characteristic impedance can rise by as much as 10 percent over the theoretical values. For microwave work most cables are nominally 50 ohms. In i-f and video work, 70-ohm cable is more generally used. U-h-f systems use both 50 and 70-ohm cable. Manufacturers are now able to hold dimensional tolerances etc., well enough so that most cables can be safely assumed to be within 3 ohms of their rated values.

Attenuation.—The attenuation in coaxial cables consists of two parts: series losses in the copper conductor, and shunt losses in the dielectric. If we assume exactly cylindrical conductors these losses can be expressed as follows:

$$A_{cu} = \frac{65.15(a+b)}{Z_0 ab}\left(\frac{\rho}{\lambda}\right)^{1/2} \text{db/cm.}$$

and

$$A_{dl} = 27.3\sqrt{k'}\,\frac{\tan\delta}{\lambda} \text{ db/cm.}$$

A_{cu} and A_{dl} are the attenuations produced by copper and insulator losses respectively, a and b are the inner and outer conductor diameters in cm., λ is the wavelength in cm., ρ is the d. c. resistivity in ohm-cm., Z_0 the characteristic impedance, k' is the dielectric constant, and $\tan\delta = k''/k'$ is the dielectric power factor (the ratio of the imaginary to the real part of the complex dielectric constant).

A word of explanation about the power factor may be helpful at this point. An imperfect (i. e. actual) dielectric may be represented by a complex dielectric constant $k_c = k' - jk''$. The angle $\delta = \tan^{-1} k''/k'$ is then the angle between k' and k_c in the ordinary complex number space. Since for any useful material, $k'' << k'$, then $\tan\delta \cong \sin\delta = \cos(90° - \delta)$, which is the power factor in the ordinary sense.

Referring back to our expression for A_{cu} and A_{dl} we see that the copper losses depend on the cable size and vary directly as the square root of the frequency. On the other hand the dielectric losses are independent of cable size and vary directly as the frequency. In the practical case the copper losses are predominant in the frequency range up to a few hundred megacycles, while the dielectric losses

become predominant in the microwave region. A numerical example for an actual case will serve as illustration:

Federal B–452–D cable (double braided PT–5)

$a=.081''$	$\tan \delta=.0009$
$b=.285''$	$Z_0=50$ ohms
$\rho=1.72=10^{-6}$ ohm-cm.	$\lambda=10.0$ cm.
$k'=2.25$	$A_{cu}=.020$ db/ft.
	$A_{al}=.111$ db/ft.

Actually the measured total attenuation was about 0.18 db/ft. leaving a discrepancy of some 0.05 db/ft. This discrepancy must be due to the greater resistance of braid over solid copper, and to geometrical factors, such as type of braid construction, and cable imperfections, such as eccentricity.

These problems cannot be treated theoretically and actual experimental work is incomplete especially in the microwave region. Work is now proceeding at Bell Telephone Laboratory on this question. Measurements made at Radiation Laboratory show that in the 3 cm. band the effect of braid is very pronounced and very troublesome. The braid not only increases the attenuation by a large amount, but the attenuation is not constant and varies in an unpredictable manner from 0.6 to 0.9 db/ft. depending on the history of the cable, the temperature, and seemingly the time of day. Work is progressing on developing usable test cables for X-band work.

In passing it may be well to mention the use of so-called Bell Laboratories "Lossy" cable. This cable, KS–8086 (70-ohm) and D–163296 (50-ohm), is widely used at Radiation Laboratory as an isolating device, especially in test equipment. The dielectric is a synthetic rubber of high power factor (about 0.006). The attenuation of this cable is around 1.2 db/ft. in the 10 cm. band. The attenuation varies nearly linearly with both the frequency and the temperature, so that its use as a calibrated attenuator is to be discouraged.

Breakdown.—For test equipment little r-f power is involved and break-down is of no consequence. However, it is often desirable to insert a flexible section in an r-f power transmission line, and this can be done easily by using cable. At microwave frequencies it is necessary to distinguish between two kinds of break-down, voltage, and thermal. True voltage break-down generally starts with corona which gradually eats its way through the dielectric, carbonizing as it goes, until flash-over occurs. Thermal break-down is directly due to the rather large attenuation in cables which causes heating of the dielectric. At high average power levels the heating can get large enough for plastic flow to occur in the dielectric until the cable shorts. For pulsed transmission we can predict that voltage break-down will

occur at high-peak powers with low-duty cycles, while thermal break-down will occur at high-average powers, i. e. high-duty cycles.

To have high-voltage break-down strength it is necessary that the dielectric be extruded tightly on the center conductor, and that all air pockets be driven out. A very simple calculation shows that if an air pocket large enough to allow corona to start exists at the inner conductor, the corona-starting voltage of the cable will be lower, and the power-handling ability of the line will be reduced by a factor of $K^{3/2}$ for the same inner and outer conductor diameters. This is based on the assumption that the maximum power which can be handled by the line is one in which the r-f voltage is just below the corona-starting potential.

Corona-starting voltages are almost impossible to measure at high frequencies. Consequently, 60-cycle measurements are taken as

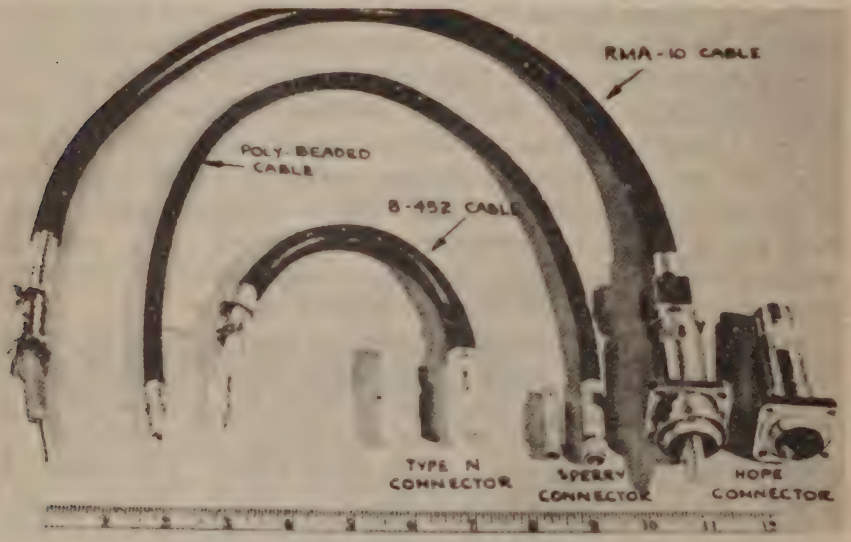


FIGURE 23.—Typical flexible cables and connectors.

indicative of the relative goodness of cables in this regard. In the cables used at Radiation Laboratory (max. size 0.350-inch diam. of dielectric) 60-cycle corona-starting potentials have been measured from 3,500 to 10,000 peak volts. At present most of our cables are well in the higher part of this range, 7,000–10,000 volts. Under r-f power in the 10-cm. band thermal failure takes place in the region near 300-kw. pulse power, 300-watts average power. While partial destruction of these cables has been observed due to corona burning, true voltage break-down has never been observed in the cable. In all cases not attributable to faulty connectors, the failure takes place at the air-dielectric interface in the connector.

Connectors.—For high power use, connectors have been developed

at Radiation Laboratory for two cables: PT–5 [0.285-inch d. o. d. (diameter of dielectric)], and for RMA–10 (0.350-inch d. o. d.). These connectors were developed to attach the cable to ⅞-inch stub-supported air-filled line. The connector has good break-down qualities because dielectric is extruded into the connector and bonded to the cable dielectric. The break-down voltage of the connector is necessarily limited by the size of the air-filled line and by the necessity of maintaining constant impedance.

For low power and test equipment use, the "type N" connector is the accepted Army and Navy standard, and is used quite generally with PT–5, RMA–16, and Western Electric "lossy" cable at Radiation Laboratory. This connector can be used up to 500 volts maximum (2.5 kw. pulse power). This connector and others are shown with typical flexible cables in figure 23.

50-Ohm Flexible Cables

Cable type	Manufacturer	i. d.	d. o. d.	o. d.	3000 mc. att. db/ft	60~ corona potential(kv.)	Rated r-f power at $\frac{1}{10}$ duty cycle (pulse kw.)	Weight (oz/ft)
		Inch	*Inch*	*Inch*				
PT–5 D. B. (B–452–D).	Federal	0.086	0.285	0.420	0.21	7.0	50–75	2.1
PT–5 (B–45)	do	.086	.285	.405	.21	7.0	50–75	1.7
RMA–10	do	.102	.371	.555	.17	9.0	75–100	3.2
RMA–10	G. E	.102	.371	.555	.24	8.0	75–100	3.2
RMA–16	Federal	.052	.186	.370	.29		[1] 2.5	1.55
RMA–16	G. E	.052	.186	.370	.33		[1] 2.5	1.7
D–163296 (lossy)	W. E	.047	.201	.315	1.2		[1] 2.5	1.8
50–ohm (lossy)	Simplex	.036	.150	.315	1.4		[1] 2.5	1.7

This table is intended to give relative average values and ratings on several cables. It should act as a guide in cable selection, but if exact values are desired it is advisable to check each cable.

[1] Assumes low-voltage connector.

A complete listing of cables and connectors meeting Army-Navy requirements as of 17 January 1944 may be found in the restricted publication "R. F. Transmission Lines for Army-Navy Use" (RE–9196B) issued by the Bureau of Ships (Code 930G).

2. COAXIAL LINES AND WAVE GUIDES

a. Propagation in Wave Guides.

The term *wave guide* ordinarily refers to hollow metal tubes through which it is possible to transmit electromagnetic energy. There are other types of wave guides, such as solid dielectric cables, through which it is possible to transmit energy but we shall not consider these types here.

That it is possible to transmit electric energy down a "one-conductor" line may be surprising to those who are accustomed to thinking

in terms of voltages and currents along two-wire transmission lines. In thinking of energy transfer in wave guides it is necessary to think of electromagnetic waves propagating down the guide and being confined by the metal walls in much the same way as commercial radio waves are confined between the earth's surface and the Kennelly-Heaviside layer. To study these waves the fundamental concepts of electric and magnetic fields must be used. Two-wire transmission lines can also be analyzed using the field concepts, but it is usually more convenient to treat them in terms of voltages and currents. With wave guides it is impossible to define a unique voltage at any point along the line.

b. Modes in General.

Solutions of Maxwell's equations subject to the boundary conditions imposed by perfectly conducting metal walls show that any given guide can propagate an infinite number of different types of electromagnetic waves. Each type, or mode, has its own individual electric and magnetic field configuration and is designated by means of it. Each mode has a lower critical frequency below which it will not be propagated down the guide. The critical frequency is called the "cut-off frequency," f_c, for that mode, and the corresponding free space wavelength is called the "cut-off wavelength," λ_c.

$$\lambda_c = \frac{c}{f_c}$$

where c is the free space velocity of light (3×10^{10} cm/sec). In a given size guide the mode with the lowest cut-off frequency is called the lowest mode for that guide and will be the only one propagated if the frequency is greater than its cut-off frequency and less than the cut-off frequency for the second lowest mode. It is a noticeable characteristic of wave guides that the cut-off wavelength of the lowest mode is comparable to the size of the guide. For example, in round air-filled guide the cutoff wavelength of the lowest mode is equal to 1.71 times the diameter, and in rectangular air-filled guide it is equal to twice the large cross-sectional dimension.

In general two types of modes are possible. The modes of one type are called "transverse electric" (*TE*) waves because their electric fields are everywhere transverse to the axis of the guide. Sometimes these waves are referred to as *H*-waves because they have components of the magnetic field parallel to the guide axis, but it is recommended that the *TE* terminology be adhered to for standardization purposes. The modes of the other type are called "transverse magnetic" (*TM*) because their magnetic fields are everywhere transverse to the guide axis. Sometimes these *TM* waves are referred to as *E*-waves because they have electric field components parallel to the guide axis.

Only a few of the lowest modes are now used in radar equipment. Since the various modes have different wavelengths in guide, it is generally impossible to make a line which will be matched for more than one mode. Also at any discontinuity, such as a junction or sharp corner, all propagating modes will tend to be excited so that it is desirable to choose a guide which will propagate only the lowest mode. Then any mode beyond cut-off will die out within a very short distance of the source. Higher modes are used only because of their special polarization or attenuation features. The wavelength in guide (λ_g) for any mode is given by

$$\lambda_g = \frac{\lambda_0}{\sqrt{K}\sqrt{1-\left(\frac{\lambda_0}{\lambda'_c}\right)^2}}$$

where λ_0 is the free space wavelength, λ'_c the cut-off wavelength for that mode, and K the dielectric constant of the material with which the guide is filled. Most guides are air-filled so that $K=1$. It should be pointed out that cut-off wavelength λ'_c also depends upon the dielectric within the guide. (See section II, 3g.) The guide wavelength for all air-filled guide is, as the above equation shows, always larger than the corresponding free space wavelength. Although this fact indicates that the phase velocity is greater than that of light in free space, energy is not being propagated at this rate, because the group velocity is always less than that of light and approaches zero as the free space wavelength approaches the cut-off value.

c. Modes in Rectangular Guides.

The lowest mode in rectangular guides is the $TE_{1,0}$ mode. An instantaneous picture of its field is given in figure 24 (a). The electric field can be represented by electric lines of force which extend from the bottom to the top of the guide. The intensity of the field varies sinusoidally along the "a" dimension, having a maximum at the middle of the guide and being zero at the edges. Along the "b" dimension it is uniform, and along the axis of the guides the intensity varies sinusoidally.

The magnetic field is pictured by means of the magnetic lines of flux. These lines always form closed loops. If the electric field is known, the magnetic field configuration can be determined. The two fields are always mutually perpendicular. For the $TE_{1,0}$ mode the magnetic lines form plane loops which are perpendicular to the electric lines.

Another useful concept to keep in mind when dealing with various modes is that of the current flow on the inside surfaces of the guide. The lines of current flow are orthogonal trajectories of the magnetic field lines at the surface of the conductor and so are readily found once

the magnetic field is known. The lines of current flow for the $TE_{1,0}$ mode in rectangular guide are shown in figure 24 (a). If an obstruction is placed in or on the inside surface of a guide in such a way as to obstruct the natural flow of the surface current, then in general a reflected wave will be caused by this discontinuity. For example,

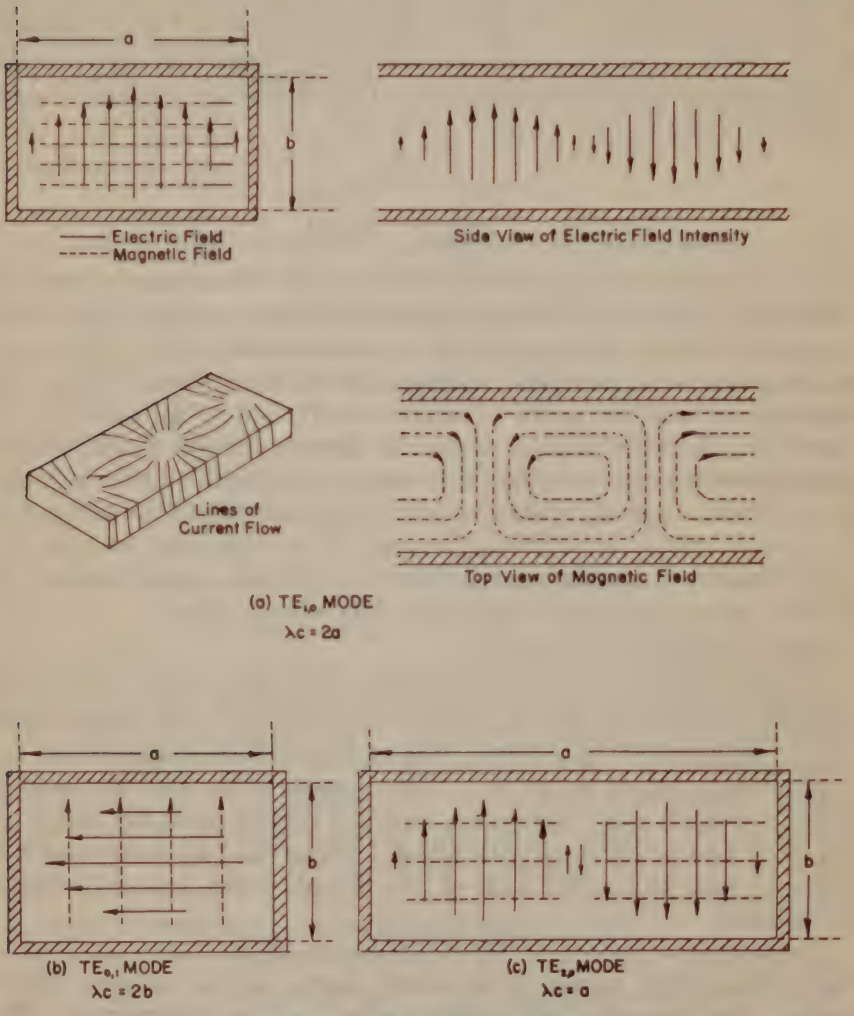


FIGURE 24.—Modes in rectangular guides.

when cutting a thin longitudinal slot in a rectangular guide carrying the $TE_{1,0}$ mode, it is necessary to cut it along the middle of the broad dimension of the guide where the lines of current flow are longitudinal. In some modes the current flow is everywhere transverse to the guide axis, in which case it is impossible to cut a longitudinal slot without badly disturbing the current flow.

The cut-off wavelength in air (λ_c) for the $TE_{1,0}$ mode is

$$\lambda_c = 2a$$

i. e., the "a" dimension of the guide must be at least one-half the free space wavelength in order that the $TE_{1,0}$ mode be propagated. In a dielectric-filled guide

$$\lambda_c' = 2a\sqrt{K}.$$

The various modes are designated by the subscripts m and n, where m and n are whole numbers. In rectangular guide the first subscript m of a $TE_{m,n}$ or $TM_{m,n}$ mode refers to the number of half cycle variations in the fields along the "a" dimension, and the second subscript n refers to the number of half cycle variations along the "b" dimension. For example, the $TE_{0,1}$ mode [see figure 24 (b)], is just like a $TE_{1,0}$ mode which has been rotated through 90°, but it has the subscript "0,1" because the fields are uniform along the "a" dimension and have a half cycle variation along the "b" dimension. The cut-off wavelength for the $TE_{0,1}$ mode is:

$$\lambda_c = 2b.$$

The next higher mode in rectangular guide is the $TE_{2,0}$ mode, shown in figure 24 (c). It is best described as two $TE_{1,0}$ modes side by side and 180° out of phase in time. The cut-off wavelength for the $TE_{2,0}$ mode is:

$$\lambda_c = a.$$

These data on the cut-off wavelengths allow us to choose a guide which will propagate only the $TE_{1,0}$ mode. The dimensions "a" and "b" of the guide must be such that

$$2a > \lambda_0 > a.$$
$$\lambda_0 > 2b$$

where λ_0 is the free space wavelength of the frequency to be propagated. The inequality

$$2a > \lambda_0$$

allows the $TE_{1,0}$ mode to be propagated; the inequality

$$\lambda_0 > a$$

prevents the $TE_{2,0}$ and higher modes; and the inequality

$$\lambda_0 > 2b$$

prevents the $TE_{0,1}$ and similar modes. The cut-off wavelength for any $TE_{m,n}$ or $TM_{m,n}$ mode in a rectangular guide is given by

$$\lambda_c = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}}, \quad \lambda'_c = \sqrt{K}\lambda_c$$

d. Modes in Circular Guides.

The lowest mode in round wave guide is the $TE_{1,1}$ mode, which is the round wave guide equivalent of the rectangular $TE_{1,0}$ mode. It is shown in cross section in figure 25 (b). Its cut-off wavelength is equal to 1.71 d, where d is the inside diameter of the guide. As a rule cut-off wavelengths for modes in circular guide cannot be expressed in easily remembered numbers, because they involve roots of Bessel functions.

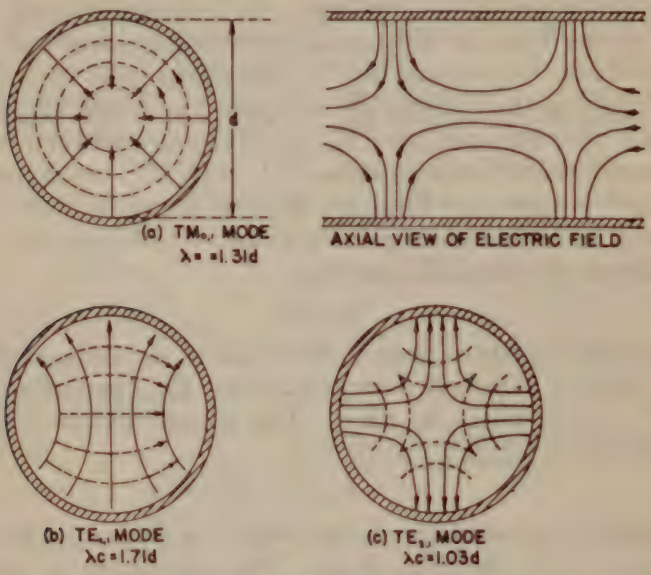


FIGURE 25.—Modes in round guides.

The next lowest mode is the $TM_{1,0}$ shown in figure 25 (a). The magnetic lines form concentric circles and lie in planes perpendicular to the guide axis. Since the magnetic lines are everywhere perpendicular to the guide axis, the surface currents must everywhere flow parallel to the guide axis. Figure 25 (a) also shows an axial view of the electric field. Each line of force emerges perpendicularly from the guide wall, turns and travels parallel to the guide axis, and then bends back into the guide.

Figure 25 (c) shows a cross-sectional view of the $TE_{2,1}$ mode which is the next higher one.

The subscripts for the circular guide modes indicate the number of angular and radial modes in the order given. Thus the $TE_{m,n}$ mode. will have m angular nodes (diameters) and $(n-1)$ radial nodes (circles).

e. Modes in Coaxial Lines.

Different modes can also exist in coaxial lines. In fact, a coaxial line can be thought of as a space between two concentric metal

cylinders. The usual mode in coaxial lines is the principal *TEM* (transverse electromagnetic) mode. This principal mode is unique in that it does not have a cut-off frequency. Any size of coaxial line can propagate it.

The wavelength for this mode is given by

$$\lambda = \lambda_0(1 - 1.7 \times 10^{-4} \alpha^2)$$

where λ_0 = wavelength in free space

α = attenuation in db per wavelength.

The difference between the wavelength of the principal mode in a coaxial line and that in free space is negligible for any reasonable attenuation. In fact

$$\alpha = 23.8 \frac{\sqrt{\rho \lambda_0}}{Z_0}\left(\frac{1}{a} + \frac{1}{b}\right)$$

is the attentuation in this mode due to conductor losses. Here:

ρ = resistivity of the conductor in ohm cm.

a, b = radii of inner and outer conductors in cm.

and λ_0 is in cm.

For a 50-ohm ⅞-inch brass line at 10 cm. wavelength, $\alpha = 0.014$ db per wavelength and the correction in wavelength is 3×10^{-7} cm.

The next higher mode appears when the mean circumference of the line is approximately equal to λ_0. Each half of this mode resembles very much the $TE_{1,0}$ mode in a rectangular guide which has been curved (see figure 26 (b)). This mode presents another limitation on coaxial lines because its wavelength is greater than that of the principal *TEM* mode. Whenever a coaxial line is large enough for the higher mode to appear, its size is comparable to that of a rectangular wave guide which can transmit the $TE_{1,0}$ mode. In the 10 cm. band the largest size of coaxial line which can be used without the danger of the next higher mode appearing is the 1⅝-inch line. In the 3 cm. band the ½-inch line is the largest permissible. Because of the limited power handling ability, attenuation and extra machining of the ½-inch line, wave guides are used almost exclusively in the 3 cm. region.

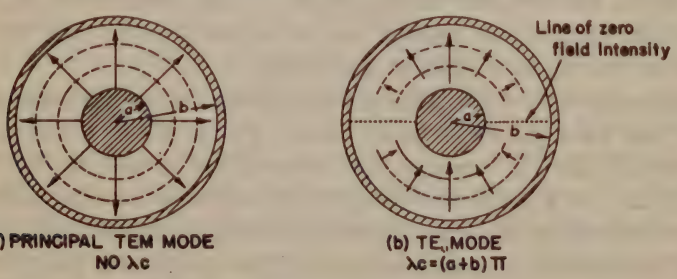


FIGURE 26.—Modes in coaxial lines.

f. Coaxial Lines Versus Wave Guides.

In choosing between coaxial lines and wave guides, the following properties should be considered: size, power-handling ability, attenuation, broad-banding considerations, ease of fabrication, and reliability in service.

Coaxial lines may be any convenient size below the point where the second mode enters. Wave guide size must lie between the limits set by the entry of the first and second modes, as discussed above. The largest permissible coaxial line is about half the size of the smallest permissible wave guide. Hence, if for a given application a wave guide is inconveniently large, a coaxial line of suitable size is the obvious choice.

The power-handling ability is greater for wave guides. This is chiefly due to their larger size. The maximum power than can be transmitted varies approximately as the cross-sectional area for any given shape of transmission line. More exactly the maximum power (P) that can be transmitted down an air-filled rectangular guide propagating the $TE_{1,0}$ mode is

$$P = E_{max}^2 \times 6.63 \times 10^{-4} ab \left(\frac{\lambda}{\lambda_g}\right)$$

where E_{max} is the maximum instantaneous electric field in the guide, a and b are the cross-sectional dimensions of the guide, and λ and λ_g are the free space and guide wavelengths, respectively.

For an air-filled coaxial line

$$P = \frac{a^2 E_0^2 Z_0}{7200}$$

where E_0 is the field at the inner conductor, a is the radius of the inner conductor, and Z_0 is the characteristic impedance of the line.

Assuming a maximum field of 30,000 volts per centimeter in each case, we see that at 10.7 cm. the 1½" x 3" waveguide will carry about 9.8 megawatts, and the 1⅝" coaxial line will carry about 4.2 megawatts The cross-sectional area of the waveguide is 24.6 cm.^{2} and that of the 1⅝" line is 9.8 cm.^{2} so that the cross-sectional average of power density is about 400 kw/cm^2 in each case. The power carrying capacity of the stub-supported coaxial line is reduced somewhat by the distortion of the normal fields at the stubs.

The attenuation of the largest permissible coaxial line is about twice that of the wave guide normally used. For instance at 10 cm. the attenuation of the 1⅝" coaxial line is about twice that of the 1½" x 3" wave guide. At a given wavelength, the attenuation of smaller coaxial line is increased by the same factor by which the diameters are reduced.

The band width of a simple coaxial line component (e. g. a quarter-wavelength shunt branch) is calculated to be greater than that for

the corresponding wave guide component by a factor of about 1.7 to 2.3 (depending on how close to cut-off the guide is operating).

The last two considerations, ease of fabrication and reliability in service, do not weigh strongly one way or the other on 10-cm. band equipment. However, as the wavelength decreases the guide becomes easier to handle, whereas the coaxial line becomes increasingly difficult to machine to the necessary tolerances and also becomes less rugged due to small size, this being especially true for the inner conductor.

Wave guides are very easy to work with for someone who has worked with coaxial lines, because exactly the same concepts of reflections, impedance, and standing wave ratios are used. If transmitting only one mode, a wave guide has a certain characteristic impedance (although its exact definition has not yet been settled), and unless the guide is terminated in it, some of the incident energy will be reflected. If measurements are made on the standing wave thus produced, the exact amount of mismatch can be found by using an impedance chart exactly as with the coaxial lines. Higher modes are present at all discontinuities in both wave guides and coaxial lines, but these higher modes are rapidly attenuated because they are beyond cut-off. The over-all effect is as if just the lowest mode were present. However, if standing wave measurements are made too near a discontinuity, the effect of the higher modes will be seen.

g. Resonant Sections.

Consider a wave traveling down a length of wave guide or coaxial line which is perfectly shorted at each end. The wave will be perfectly reflected at each end and will continue to be reflected for a larger number of times, depending upon the attenuation encountered. The standing wave pattern thus produced will depend upon the exact manner in which the reflections combine. If the length of the line is an integral number of half wavelengths, the reflections will all add in phase, and the amplitude of the standing wave will be many times that of the original wave. In this case the length of coaxial line or wave guide is said to be "resonant." At any given frequency the degree of resonance is a rapidly changing function of the exact length, and conversely, for any given length the degree of resonance is a rapidly changing function of frequency. Since these sections of coaxial line and wave guide are so frequency sensitive, they can be utilized as wave meters of high resolution. The section of wave guide can also be though of as a *cavity resonator*, i. e., a hollow metal cavity which can support certain modes of oscillation, the wavelengths of which are a critical function of the geometry of the cavity. Conversely, it is sometimes convenient to think of cavity resonators as wave guides of resonant length.

3. WHAT CAN BE DONE WITH WAVE GUIDES

a. Obstacles in Wave Guides

Discontinuities of any sort in a wave guide may produce undesirable reflections which must be removed in some way. Obstacles may be placed in the guide in such a manner as to produce reflections which cancel these. It is convenient to describe the effects produced by an obstacle placed in the guide in terms of impedance or admittance in analogy to transmission line theory.

The matching devices most commonly used in systems are diaphragms and tuning screws. The diaphragm, or window, consists of an aperture in a thin piece of metal, placed transversely across the wave guide. Such a window, provided the metal is thin and perfectly conducting, has a pure shunt susceptance which may be inductive or capacitive, depending upon the orientation of the slit. For an inductive window, the edges of the slit are parallel to the electric field in the lowest mode in the guide. Figure 27 gives theoretical and experimental values of the shunt susceptances of such windows, and the expressions from which these quantities may be calculated. The fact that the experimental values lie slightly above the theoretical is explained by the finite thickness of the diaphragm which does not enter into the formulas.

The capacitive window has the edges of the slit perpendicular to the electric vector. For experimental values of the susceptance of this kind of window, see figure 28. It is to be noted that here the thickness of the diaphragm plays such an important role that this quantity must be included in the theoretical expression used to calculate the susceptance. When the r–f plumbing for a system is designed, mismatches of various sorts are corrected by means of diaphragms which are soldered in place and are not variable in any way. Capacitive windows are of no use where large amounts of power are involved, as they break down. In certain cases, it may become necessary, because of elements which fail and must be replaced, to have a variable reactance somewhere in the line so that retuning is possible. In such cases, tuning screws are used; i. e., small cylindrical posts projecting into the broad face of the guide affording capacitive susceptance [2] which varies with the amount of penetration of the post into the guide. It is sometimes possible to have but one screw in a fixed position provided that electrical distances are preserved, but in general, three screws are provided $\lambda_g/4$ apart along the guide. By using either one or the other adjacent pair, it is possible to match any element to the guide. Where high powers are used, there is danger of break-down at

[2] As the penetration of the screw into guide increases, the capacity susceptance increases up to the resonant point, where the susceptance becomes inductive. In actual practice, the screws are never used inductively.

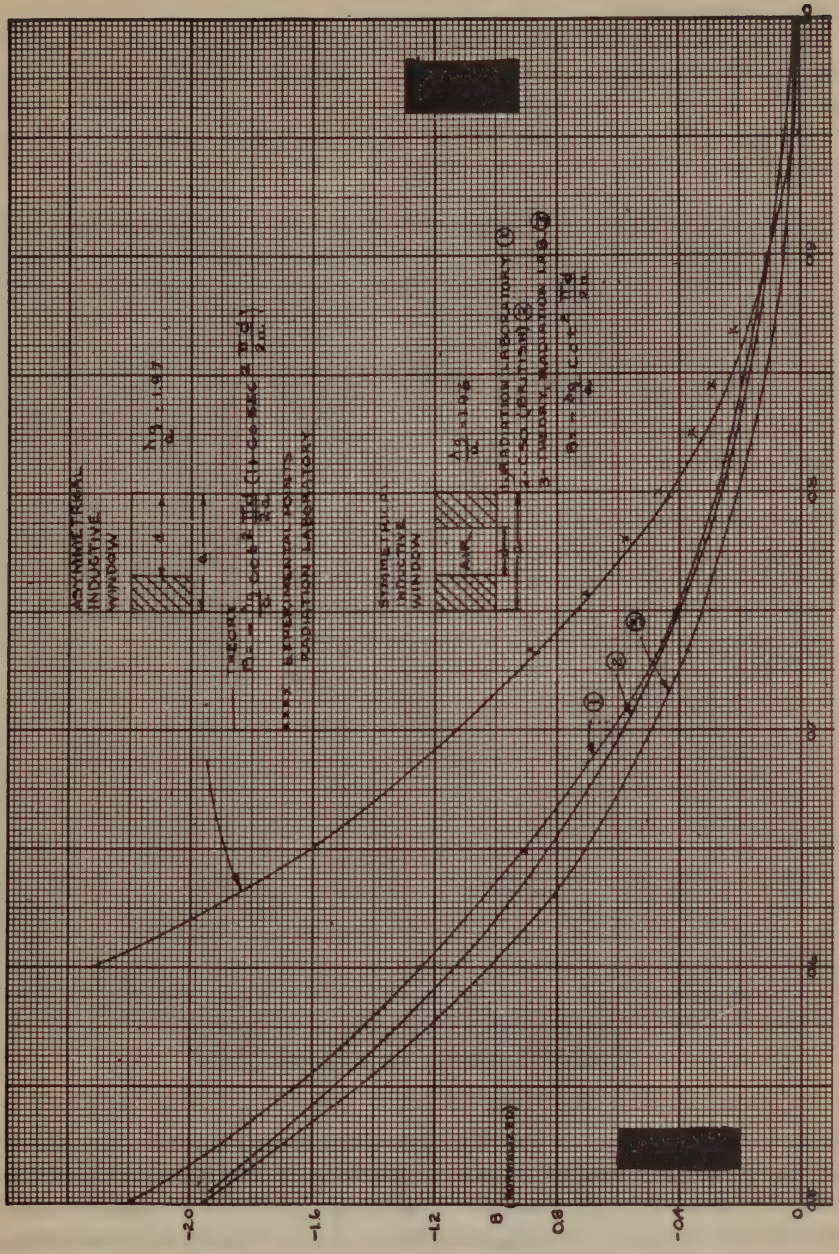


FIGURE 27.—Susceptance of asymmetrical and symmetrical inductive windows.

the screws and this limits the usefulness of screws for any but very small penetrations.

It is sometimes desired to insert into the wave guide some element which will either transmit a desired mode and no other or reflect completely the power in a certain mode. For this purpose resonant structures are used. A thin metal rectangular ring of the proper dimensions placed across the rectangular guide (supported by dielectric) will

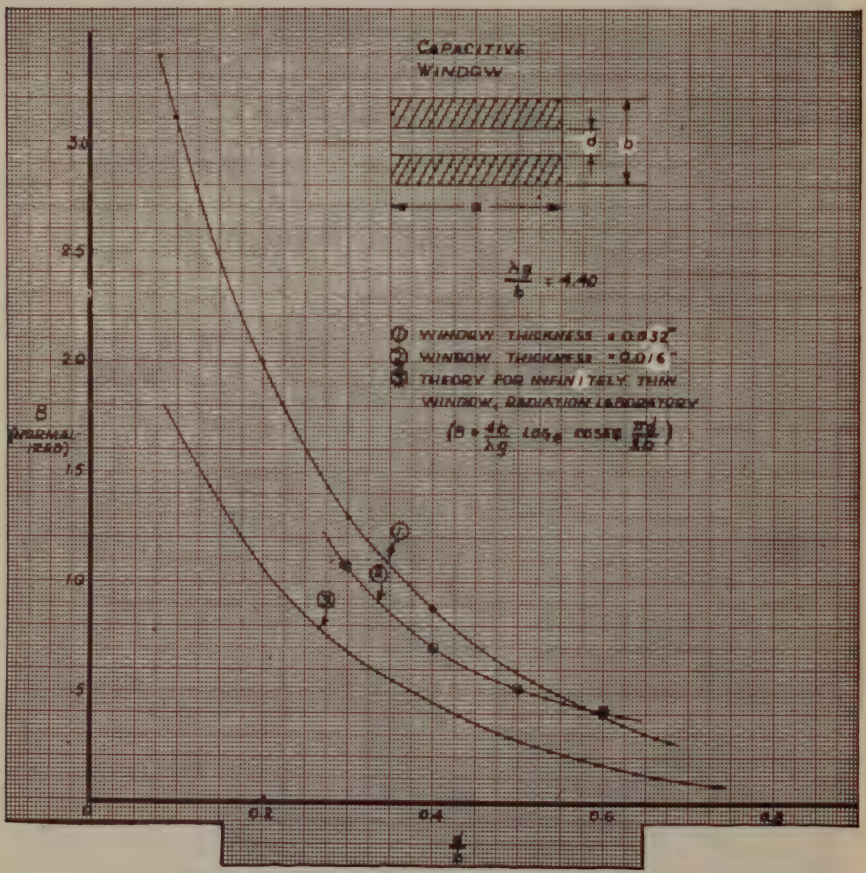


FIGURE 28.—Susceptance of a symmetrical capacitive window.

reflect, for example, all the $TE_{0,1}$ mode incident upon it; an opening of the proper dimension in a thin metal diaphragm across the guide will transmit all such power. In circular guides, a circular resonant ring or slit is used. Resonant slits are also of value for transmitting power at low level but rejecting it at high level, the slit being so narrow that break-down occurs and the resonant condition is spoiled. Thus such a device may be used as a TR box or as an anti-TR. The so-called Beetle TR used with a tube mixer is of this type.

b. Bends and Twists in Wave Guide.

It is often necessary to employ considerable lengths of wave guide, and very seldom does it occur that this guide is able to proceed in a straight line or even in one plane. Hence, bends and twisted sections must be used and, in some instances, flexible guide. Bends may be made in a rectangular guide either in the *H*-plane or in the *E*-plane. They may be smooth, gradual, bends, or right angles. In the latter case, the outer corner of the guide is cut off at a 45° angle. (See figure 29.)

Smooth bends in rectangular guides are theoretically almost perfect in that the reflections caused by them are very small. Actually, however, it is difficult to make a bend in such a way that the

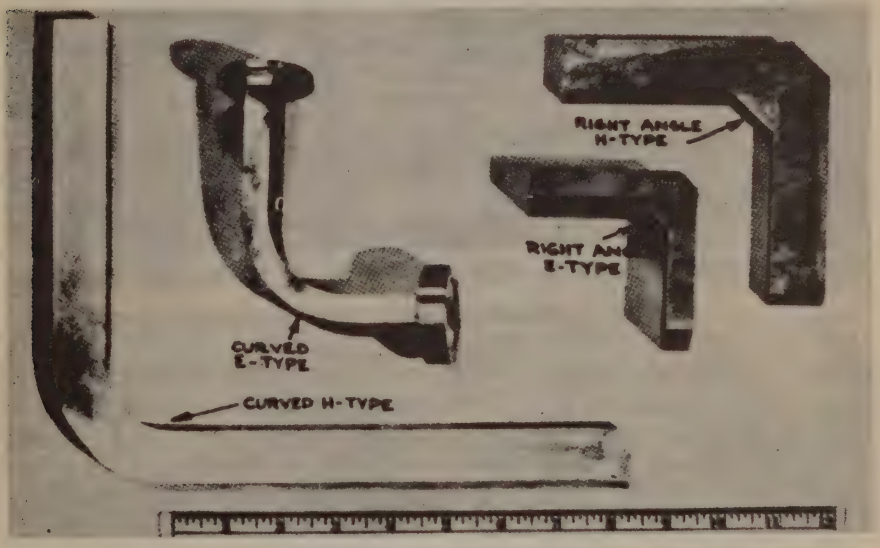


FIGURE 29.—Corners in wave guide.

guide retains its proper width and that discontinuities are avoided. By taking great care, it is possible to have a reflection less than that corresponding to a standing wave ratio of 1.05 in power if the radius of curvature of the bend is about one wavelength (in guide) or more. A wave guide may be safely twisted (such that the direction of the electric field is rotated through 90°) provided, again, that the guide is not deformed unintentionally in the process. The length of the twist should be at least of the order of one guide-wavelength. See examples in figure 30.

Bends in guides of circular cross sections are more likely to cause trouble as the polarization is not defined by the shape of the guide and the radiation is made elliptical in going around a bend. This effect may be compensated in any given system by squeezing the

round guide in such a way as to remove the ellipticity, or, in the event that there is more than one bend, they may cancel each other.

Flexible guide may be used with the understanding that the attenuation is somewhat higher than in solid guide, and some reflection will

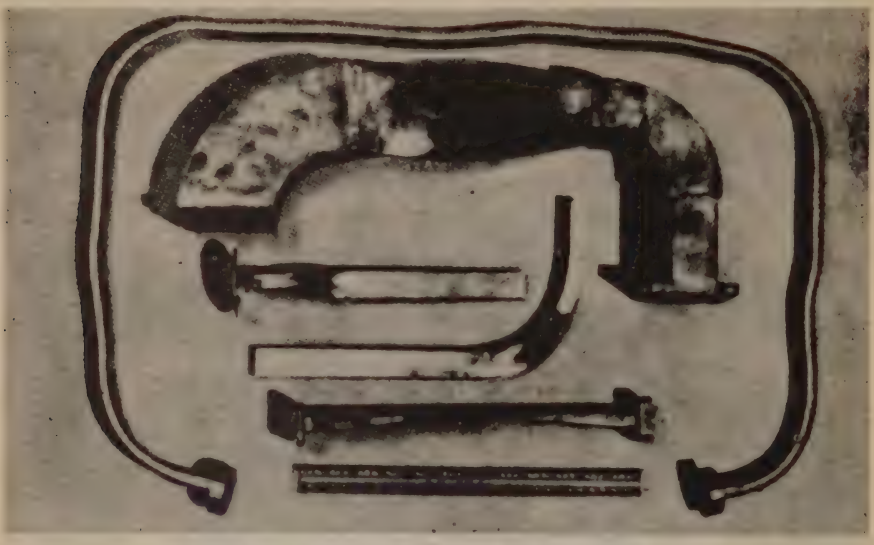


FIGURE 30.—Twisted, curved, tapered, and flexible wave guide.

occur. Its use in the field is to be discouraged inasmuch as the attenuation seems to increase markedly with time, and losses as high as one db. per foot have been measured in old samples. Sparking is also observed on high power.

c. Joints in Wave Guide.

The losses and reflections from a joint between two sections of tubing are negligible if care is taken so that the sections line up and touch. It is considered good design practice to use the more complicated "choke" connector in service applications because clean, flat, parallel surfaces are hard to achieve and maintain.

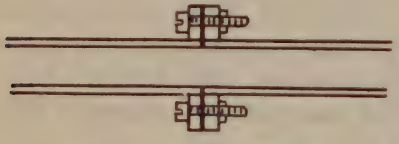


FIGURE 31.—Flange-to-flange joint.

A very simple joint is made of flat flanges which are soldered on to the ends of the tubing and bolted together. (See figure 31.) Tests have shown that this joint can be made better than even the average soldered guide-to-guide joint if care is taken to make uniform contact.

A more complicated joint, which is very common because of its flexibility, is the choke-flange joint, as shown in figures 32 and 34.

The L-shaped cavity may be thought of as a one-half wavelength transmission line into which flows the conduction current in the walls of the guide. The current and voltage amplitude distributions along a shorted two-wire transmission line are as indicated in figure 33.

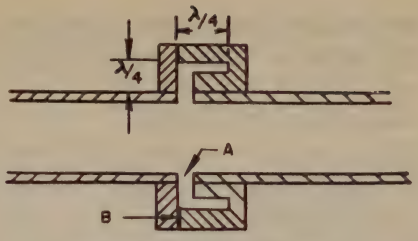


FIGURE 32.—Choke-flange joint

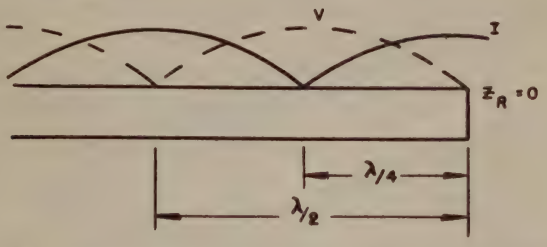


FIGURE 33.—Voltage and current distribution along a shorted transmission line.

Since gap *A* in figure 32 is ½λ from the shorted end of the L, the voltage across the gap is zero while the current is equal to the conduction current in the guide walls. On the other hand, point *B* occurs

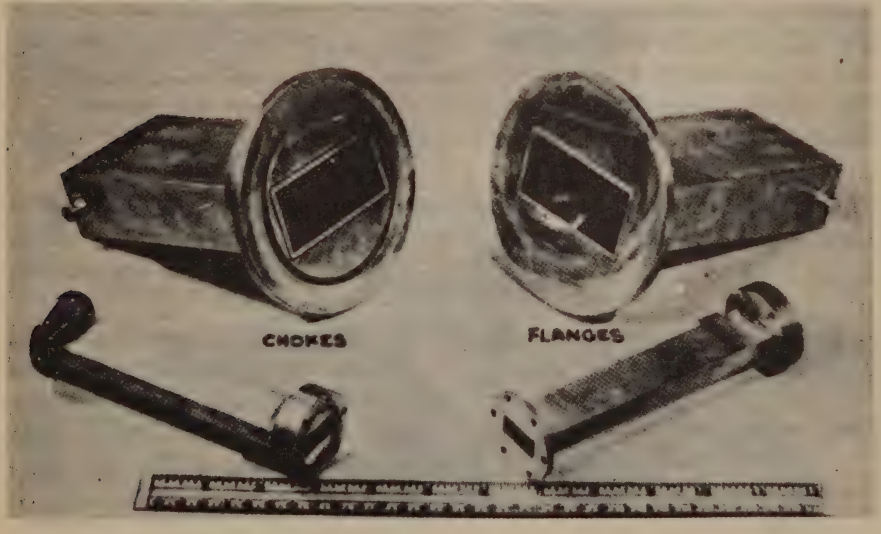


FIGURE 34.—Choke-flange joints in 10 and 3 cm. wave guide.

at the ¼λ point where the current is zero. Hence the contact at *B* need not be low resistance.

As a matter of fact, such choke couplings are often used as a "wobbly" or nonrigid connection between two wave guides. In this

application there is no contact at B. Instead a gap of about $\frac{1}{16}\lambda$ is left between choke and flange. Then for motions of about $\frac{1}{16}\lambda$ in any direction, the standing wave ratio is below $r=1.3$ and power leakage is below 0.3 db.

As a rigid connection, the loss in a choke-flange joint is of the order of 0.01 to 0.03 db. The loss in a well-made contact joint is about 0.05 db.

d. Plungers.

It will be remembered that for the $TE_{1,0}$ mode the conduction currents in the walls of rectangular guide flow from the roof across the face of a shorting plate to the bottom of the guide. (See figure 35.)

FIGURE 35.—Current flow on shorted guide.

In making plungers (i. e. movable shorting plates) care must be taken not to interrupt this flow of current in different ways at different positions of the plunger, if erratic behavior is to be avoided and if the position of the voltage node relative to the plunger is always to be the same. Furthermore, measures must be taken to insure that power does not leak past the termination.

With a cup, contact is made with the walls of the guide ¼ wavelength from the plate; that is, where the longitudinal flow of current in the walls of the guide would be zero. Losses are about 0.08 db (figure 36).

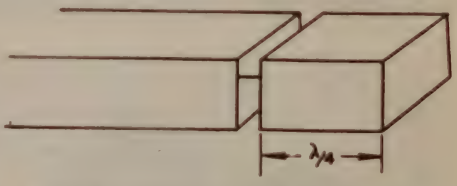


FIGURE 36.—Quarter wavelength cup termination.

Choke plungers employ ½-wavelength cavities similar to those in choke flange joints, as illustrated in figure 37. No actual mechanical contact is made with the walls of the guide at A in the front face of the plunger. Contact is made at B where the current in the cavity, considered as a ½-wave transmission line, is zero. The choke plunger appear to be the most satisfactory type, with losses of 0.02 db.

FIGURE 37.—Choke plunger.

Practical plungers are shown in figure 38.

e. T-joints in Rectangular Guide.

T-joints in rectangular guide have close analogies to side stubs in two-wire transmission lines. The side arms can be thought of as being

either in series or in parallel with the main line depending upon their construction. Side arms which lie in the electric plane, i. e., branching from the wide dimension of the guide, are in series with the main line.

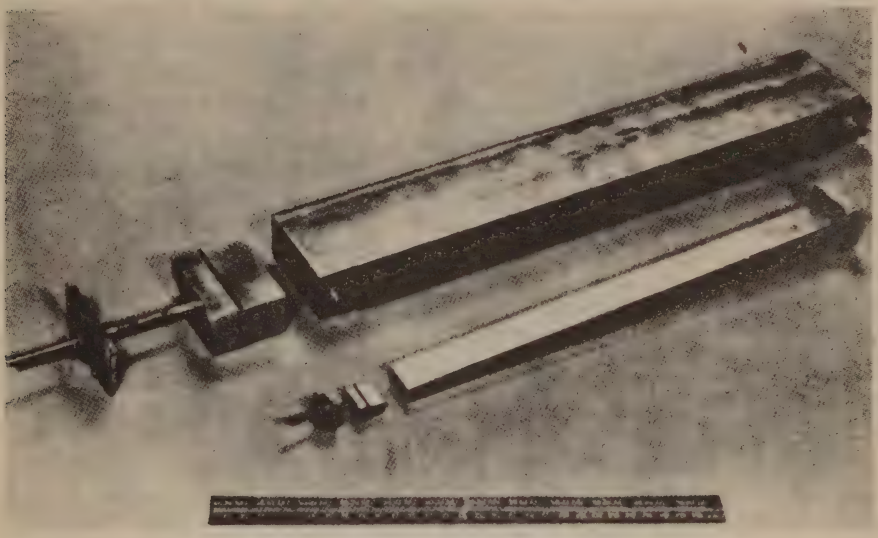


FIGURE 38.—Plungers in 10 and 3 cm. wave guide.

Side arms which branch in the magnetic plane, i. e., from the narrow dimension of the wave guide, are in shunt with the main line. (See figure 39.)

Side arms in series (*E-plane*).—Let *A* and *B* in figure 40 (A) represent the two symmetrical arms of the T-joint, and *C* the perpen-

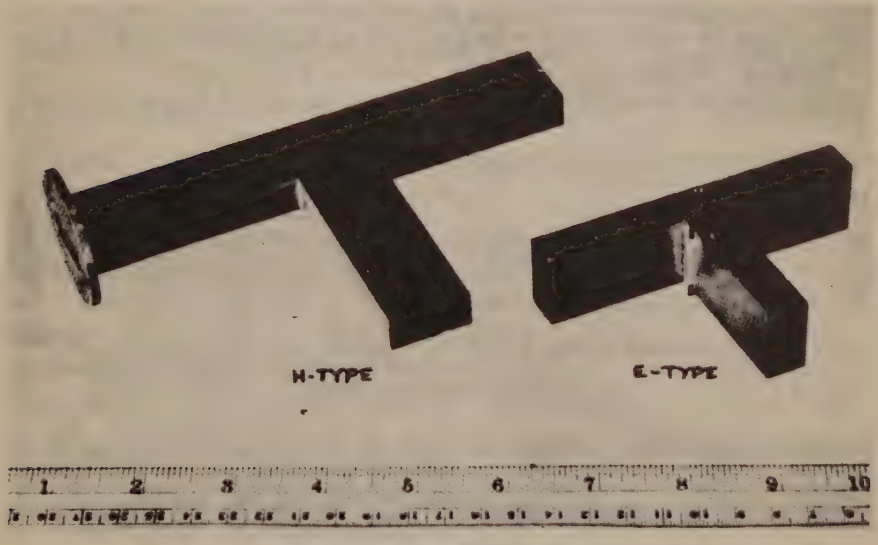


FIGURE 39.—Two types of T's in wave guide.

dicular side arm. If the plunger in C is placed slightly less than $\lambda_g/4$ from the opening, all of the power coming down arm A will be reflected. The standing waves set up in A will have a voltage minimum nearly $\lambda_g/4$ from the midpoint of the opening. This standing

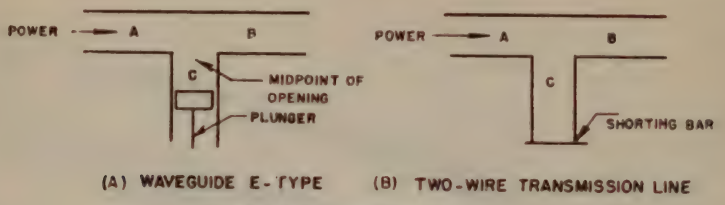


FIGURE 40.—Side arms in series.

wave pattern is identical to that produced on the two-wire transmission line shown in figure 40 (B) if the shorted side stub is a quarter wavelength long. The impedance of the stub at the junction appears to be infinite so that all the power is reflected and a voltage maximum appears. (See figure 41.)

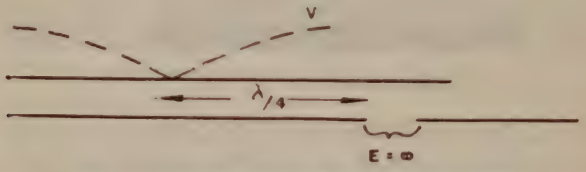


FIGURE 41.—Voltage distribution along an open-ended line.

It is also possible to have perfect straight-through transmission, i. e., from A to B, if the plunger is effectively $\lambda_g/2$ from the opening. In this case, the side arm presents a zero series impedance at the opening. The actual plunger distance is $0.45\lambda_g$, i. e., slightly less than $\lambda_g/2$ due to fringing effects.

Side arms in parallel (H-plane).—If the plunger in figure 42 (A) is

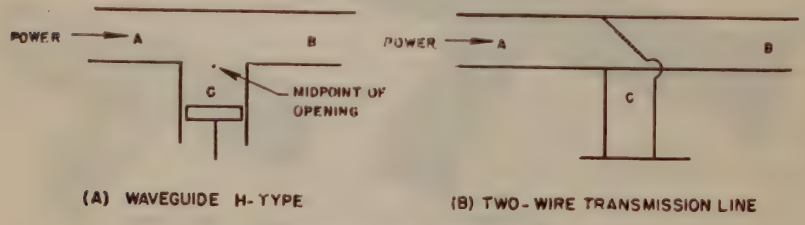


FIGURE 42.—Side arms in parallel.

placed an effective $\lambda_g/2$ from the opening, all the power coming down arm A will be reflected at the opening, and a standing wave pattern will be set up which appears to have a voltage minimum very nearly at the

midpoint of the opening. The actual plunger position is again less than $\lambda_g/2$ due to fringing effects. This pattern is identical with that on a two-wire transmission line shunted by a shorted stub which is a half wavelength long. (See figure 43.)

There is also a position of the plunger in C for which it is possible to transmit power from A to B without any reflection from the junction. In this case the plunger distance is $\lambda_g/4$, if we again neglect fringing. It is to be noticed that although the effects of the side arms in the E and H-planes are as if they were in series and shunt, their electrical lengths are different from those of the analagous side stubs in two-wire transmission lines, due to the fringing effects. This is especially true in the branching from the narrow side. (See Table 1, page 36.)

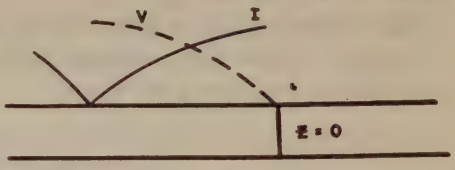


FIGURE 43.—Voltage and current distributions along a shorted line.

Plunger in B.—With the plunger in the B arm it is possible to have perfect reflection at the junction if the plunger distance l is chosen correctly. For the E-type joint, this distance is $\lambda_g/4$. and for the H-type, it is $\lambda_g/2$.

However, for no position of the plunger is it possible to have perfect transmission around the corner from A to C (figure 44). The minimum standing wave ratio in voltage obtainable in A by moving the plunger in B is about 1.3 for the E-type joint and about 1.6 for the H-type. Therefore, in order to match from A to C, it is necessary to put a matching diaphragm in one arm, say C (figure 46).

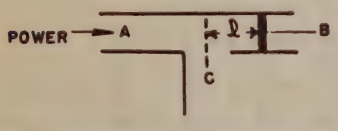


FIGURE 44.—A to C transmission in wave guide.

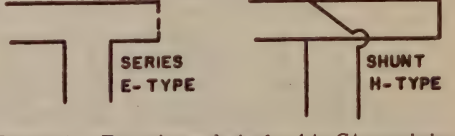


FIGURE 45.—Two wire analogies for A to C transmission in wave guide.

For any position of the plunger in B, it is possible to find a diaphragm which will allow perfect transmission from A to C. The diaphragm in the T's now in use were calculated to give a perfect match with the plunger in the position which gave the lowest standing wave ratio in the unmatched T. Inductive windows (in the narrow dimension) were chosen for both E and H sections because they are less liable to cause breakdown than capacitive windows.

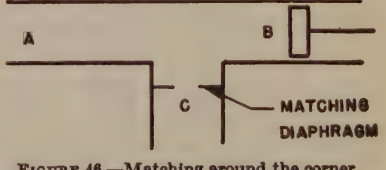


FIGURE 46.—Matching around the corner.

If the power is incident at C, the results are exactly the same as when the power is incident at A. There is total reflection, partial transmission for the unmatched joint, and perfect transmission with the diaphragm for exactly the same plunger positions as when the power is incident at A.

The following table summarizes the results for the case of 3 cm. T-joints. These experimental data agree very well with theoretical expressions developed by Frank.

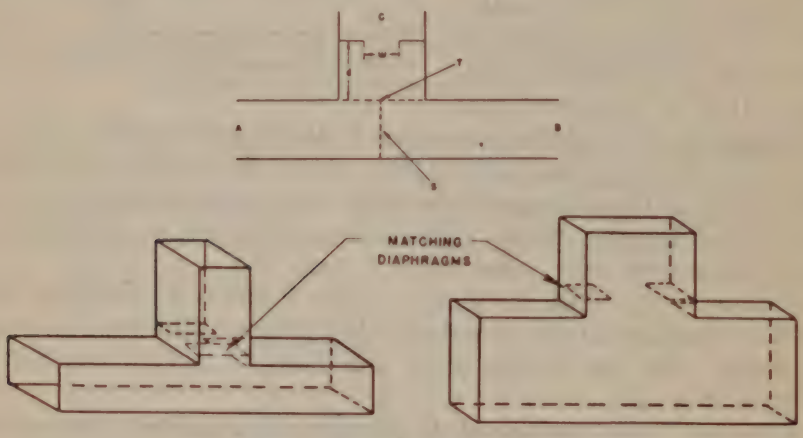


FIGURE 47.—General schematic—T joints.

TABLE I

Unmatched T's. (no matching diaphragms)

Type of T	Transmission	Plunger in	r_{min}	L_0/λ_g	B_0	L_0'/λ_g	B_0'
E	$A \rightarrow B$	C	1.0	0.45	0.02	0.22	0.27
E	$A \rightarrow C$	B	1.3	.50	.02	.26	.27
E	$C \rightarrow A$	B	1.3	.50	.23	.26	.22
H	$A \rightarrow B$	C	1.0	.57	.02	.35	.00
H	$A \rightarrow C$	B	1.6	.69	.24	.46	.00
H	$C \rightarrow A$	B	1.6	.69	.37	.46	.36

TABLE II

Matched T's. (with matching diaphragms)

Type of T	Transmission	Plunger in	r_{min}	L_0/λ_g	B_0	L_0'/λ_g	B_0'
E	$A \rightarrow B$	C	1.0	0.48	0.04	0.22	0.27
E	$A \rightarrow C$	B	1.1*	.52	.02	.26	.28
E	$C \rightarrow A$	B	1.0	.52	.23	.26	.23
H	$A \rightarrow B$	C	1.0	.62	.45	.41	.46
H	$A \rightarrow C$	B	1.0	.68	.49	.46	.49
H	$C \rightarrow A$	B	1.0	.68	.20	.47	.40

*The difference between this value and the theoretical value of 1.0 is probably due to experimental error.

r_{min}=minimum standing wave ratio in voltage obtained by plunger in C or B.

L_0/λ_g=corresponding plunger distance, measured from line t when in c and from line s when in b.

L_0'/λ_g=plunger distance for complete reflection.

B_0=distance of minimum point from line s or t in guide wavelengths for minimum reflection.

B_0'=same for complete reflection.

Matching diaphragm dimensions: (for λ=3.20 cm.) (See fig. 17.)

E-type: w=1.84 cm. d=0.68 cm.
H-type: w=1.58 cm. d=1.07 cm.

Frequency sensitivity.—Table III shows some estimates of the frequency sensitivity of E- and H- plane T sections at a wavelength of 3.2 cm. It is assumed in each case that the section is perfectly matched with the plunger at the corresponding L_0 position (see preceding table). Column A shows the estimated standing wave ratio for a displacement of the plunger of ± 1 mm. Column B shows the estimated ratio for a wave length change of ± 2 percent. The figures in column B have not been checked experimentally.

TABLE III

Type of T	Transmission	Matched	A r for $\Delta L=\pm 1$ mm.	B r for $\Delta\lambda=\pm 2\%$
E	$A\rightarrow B$	no	1.12	1.05
E	$A\rightarrow B$	yes	1.12	1.05
E	$C\rightarrow A$	yes	1.15	1.06
H	$A\rightarrow B$	no	1.08	1.04
H	$A\rightarrow B$	yes	1.18	1.10
H	$C\rightarrow A$	yes	1.11	1.02

The above tables are results obtained on ½-inch x 1 inch wave guide. Since the 10 cm. 1½ inches x 3 inches wave guide (outside dimensions) is not quite a perfect scale up, these results cannot be exact for this band. Equivalent 10 cm. data are not available, but all measurements made in this region have agreed with Frank's theoretical results to within experimental error.

f. Effects of Dirt, Water Condensation, and Silver Plating.

As mentioned above, the loss of power which takes place upon passage through a long line is primarily caused by the ordinary I^2R, r-f losses of the currents flowing in the walls of the line. These currents flow in the surfaces down to a depth depending upon the frequency of the radiation and the conductivity of the wall material and the losses are, at constant λ, simply proportional to this depth, which

is called the "skin depth". The skin depth, wavelength, and resistivity are related by the expression:

$$\delta = 0.029\sqrt{\frac{\lambda\rho}{\mu}}$$

where δ is the skin depth in cm., λ the wave length in cm., ρ the resistivity in ohm-cm., and μ the relative permeability.

One might conclude from this that the condition of the surface of the line would be very important in determining the losses. This is fortunately not the case. The metallic oxides and carbonates which form on the surface through the corrosion processes have a very small conductivity and hence a large skin depth. The depth in most cases is large compared to the thickness of the coating itself.

The losses are, therefore, little affected by corrosion. As an example of the magnitude of the extra losses to be expected, some measurements have been made on the effect of anodizing aluminum wave guide used in the 3 cm. band. The anodizing process is an electrolytic one in which a coating of aluminum oxide is formed on the surface of material. The calculated loss in the untreated guide was 0.03 db/ft., and the measured value agreed well with this. Anodizing increased the loss to between 0.04 and 0.05 db/ft. or about a 50 percent increase. An additional coating of weatherproof paint caused a further increase to 0.06 db/ft., or double that of the bare metal.

The introducing of dielectric, nonconducting materials will also increase the loss, but again the effect is small, chiefly because so little material is generally added. An important exception, however, is the case of water. Small amounts of moisture condensed on the interior surface of an r–f line cause losses up to perhaps 0.5 db./ft in 3 cm. band wave guide for example. This is caused not only by the large dielectric loss which is characteristic of water but also because of the enormous dielectric constant of water. We may say, roughly speaking, that all the power flows through the water even though it be a thin film, and hence the losses are high. Fortunately, water is almost the only substance likely to be encountered which produces such high losses.

g. Effect of Dielectric.

The effect of an ideal lossless dielectric in a wave guide is to increase the apparent linear size of the guide by a factor $\sqrt{K}$, where K is the dielectric constant of the material added. The new cut-off wavelength (λ'_c) for a given mode is given by

$$\lambda'_c = \sqrt{K}\lambda_c$$

where λ_c is the cut-off wavelength in the empty guide. The space variation of the electric and magnetic fields in any transverse section

is identical with that of the mode in empty guide. The wavelength, however, is shorter than that in empty guide, corresponding to smaller phase and group velocities. The specific wave impedance of the guide is also changed by the addition of a dielectric, always being decreased for *TE* modes and usually being decreased for *TM* modes. The *TM* case is more complicated that the *TE* case. Under the two conditions that

$$(\lambda/\lambda_c)^2 > \tfrac{1}{2}$$

and

$$K < \frac{(\lambda/\lambda_c)^2}{1-(\lambda/\lambda_c)^2}$$

the specific wave impedance for a *TM* mode in a dielectric-filled guide will be greater than in an air-filled guide. The losses of a guide filled with a non-ideal dielectric are greater than those of an empty guide due to absorption of energy by the dielectric.

If the dielectric constant of an actual dielectric is given by

$$K = K' - jK'',$$

then the loss length L per decibel for guide filled with that dielectric is given very closely by (providing we are not operating too near cut-off in the dielectric-filled guide)

$$L = 0.0367 \frac{\lambda_0}{K''} \sqrt{K' - \left(\frac{\lambda_0}{\lambda_c}\right)^2}$$

where λ_0 is the free space wavelength, and λ_c the cut-off wavelength for similar air-filled guide. For polystyrene, as an example ($K' = 2.56$ and 2.43, and $K'' = .00205$ and .00234 at 10 cm. and 3 cm. respectively), $L = 2.6$ meters per db, and 0.70 meter per db at 10 cm. and 3 cm. respectively, for the usual wave guide sizes.

The fact that the presence of a dielectric decreases the guide wavelength is utilized by one form of line stretcher which inserts controllable amounts of dielectric into the guide.

h. Wave Guide Transformers.

Transformation problems which occur whenever it is desired to change from one wave guide to another are generally of two types.

(*1*) *Lowest mode to lowest mode.*—In one type the change is from the lowest mode in one guide to the lowest mode in another. This change is accomplished by methods analogous to those used in coaxial line technique, namely, through the use of tapered sections and quarter-wavelength transformers. The most common change of this type is from the $TE_{1,0}$ mode in rectangular guide to the $TE_{1,1}$ mode in circular. Tapered sections are used for this purpose, because a taper can include a gradual transformation of shape as well as of size. (See figure 48.)

When a change of shape is not required, as is the case when changing from one size of rectangular guide to a different size of rectangular guide, the wave guide analogy of the coaxial quarter-wavelength transformer can be used. The wave guide picture, however, is not as simple as that for coaxial lines, for at an abrupt junction between two different sizes of guides, one not only sees the impedance of the next guide but also a reactive window in parallel with it. This is known as the *junction effect* and must be taken into account when designing the quarter-wavelength transformer. Also just which characteristic impedance to use depends upon whether the difference

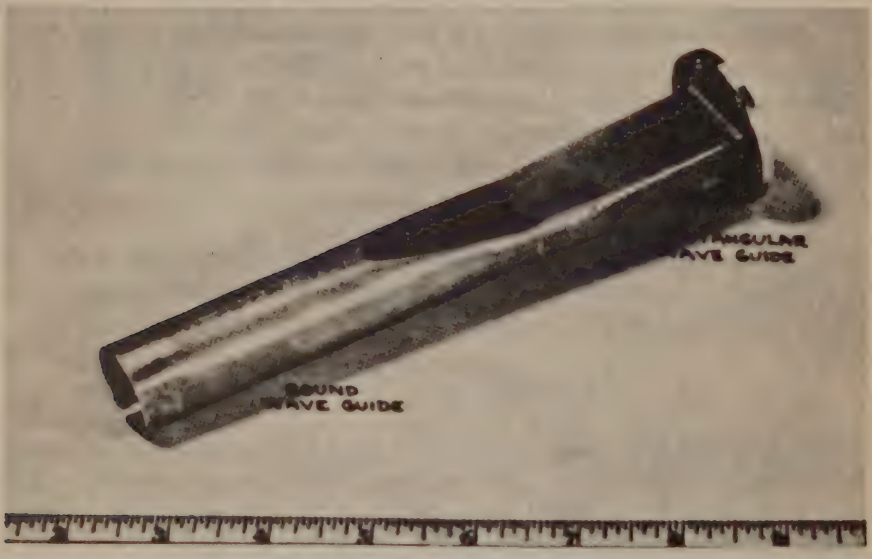


FIGURE 48.—$TE_{1,1}$—$TE_{1,0}$ transformer.

between the two guides being joined is in the "a" (broad) dimension of the "b" (narrow) dimension or both.

The transition from one dielectric to another in the same size guide can be handled with quarter wavelength transformers without any junction effect to be considered, for the electromagnetic fields are not distorted at this junction. For example, the matching from a medium of dielectric constant K_1 to one of dielectric constant K_2 could be accomplished by inserting a quarter wavelength of a third medium whose dielectric constant is the geometric mean of the other two. More often, however, the transition is accomplished by a simple tapering of the dielectric materials. This latter method is easier mechanically, less frequency sensitive, and avoids the introduction of the intermediate dielectric.

The guiding principle in the determination of the length of the transition is the fact that the reflection coefficient looking at the input

of the transformer varies with the length of the transformer as shown in figure 49. One notes from this figure that the reflection from a tapered section decreases in an oscillating manner with the length of the transition. Although a long transition will always give low reflection, a short one may be just as good (if not better), if properly selected. However, a long taper will generally be less frequency sensitive.

(2) *Lowest mode to higher mode.*—The second type of transformation problem is that of changing from the lowest mode in one guide to a higher mode in another. The basic idea to keep in mind is that if a junction is formed between two guides, one of which can support more than one mode, then in general all of the propagating modes

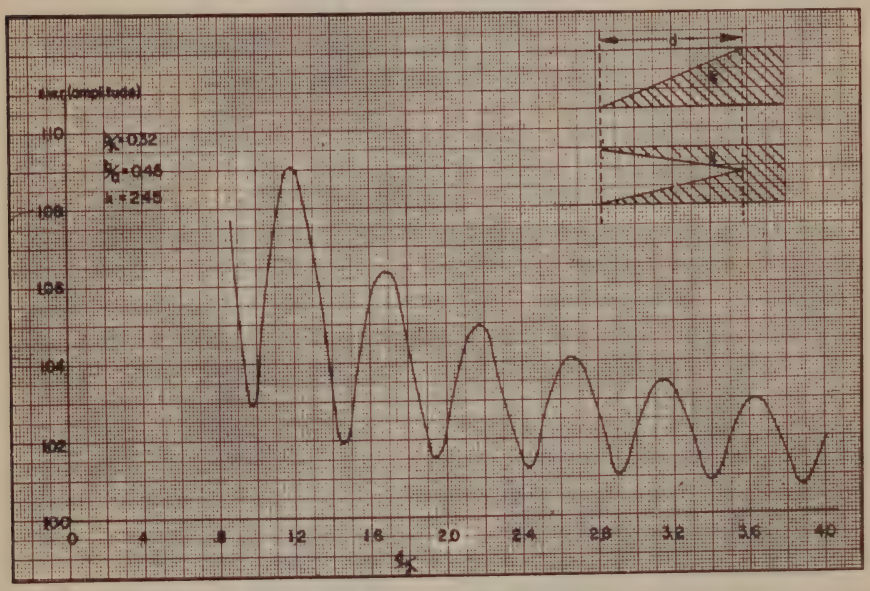


FIGURE 49.—Reflections from a tapered section.

will be set up at the junction. The design of such a junction is concerned with the attempt to fix its geometry by symmetry considerations so that only the desired mode is excited, or at least so that it is preferentially excited. Although preferential excitation can generally be accomplished, the unwanted modes can, nevertheless, be troublesome if there are certain lengths in the higher mode guide which correspond to unwanted mode resonances. These resonances must be circumvented either by correct choice of lengths or by introduction of filters for the unwanted modes (resonant rings, dipoles, etc.). The filters do not destroy the resonances but limit them to narrow regions. The type of excitation or transformation must also be chosen so as to be low Q (i. e. not frequency sensitive) and well matched. The final matching may be accomplished by an

inductive window near the junction. Transformers have been designed which are satisfactory over a 2% band.

The most common transformation of the second type is from $TE_{1,0}$ in rectangular guide to $TM_{0,1}$ in round. Two kinds of coupling are used: (1) wave guide coupling and (2) dipole coupling. As an example of wave guide coupling consider the stub and ring transformer shown in figure 51 (A). It is also shown in outline in figure 50.

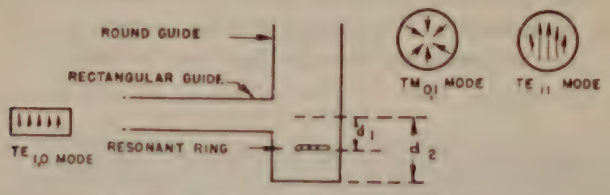


FIGURE 50.—Stub and ring $TE_{1,0}$—$TM_{0,1}$ transformer.

The stub is effectively in series with the rectangular and round guides. The resonant ring will completely reflect all of the $TE_{1,1}$ mode incident upon it and will not seriously effect the $TM_{0,1}$ mode. Therefore in this particular type of transformer the ring is placed at a distance d_1 from the junction, where d_1 is a quarter-wavelength for the $TE_{1,1}$ mode. Then for the $TE_{1,1}$ mode the stub represents an infinite impedance in series with the rectangular and round guides. The distance d_2 is chosen as to be a half-wavelength for the $TM_{0,1}$ mode so that for this mode the stub represents a very small impedance. Thus the $TE_{1,1}$ mode is suppressed and the $TM_{0,1}$ encouraged. The resonant ring can be supported by a dielectric cylinder. (See figure 51 (A).) However, this dielectric support is objectionable for field use. Another solution to the problem is the stub transformer shown in figure 51 (B). The diameter D_1 of the stub is chosen so that the ratio of the guide wavelength of the $TE_{1,1}$ mode to that of the $TM_{0,1}$ mode is approximately 2:3. Then it is possible to choose the stub length so that it is effectively a $\frac{3}{4}\ \lambda_g$ stub (high impedance) for the $TE_{1,1}$ mode and a $\frac{1}{2}\lambda_g$ stub (low impedance) for the $TM_{0,1}$ mode. Thus the same effect is achieved as was had with the stub and ring transformer. The diameter D_2 is made larger than D_1 to provide minimum attenuation. A matching diaphragm is necessary in this transformer. The two-stub transformer (figure 52 (B)) is essentially the same as the the stub transformer (figure 51 (B)) with the rectangular wave guide stub replacing the matching diaphragm.

The coaxial-stub transformer (figure 52 (A)) is of the same general type as those discussed above. The only difference is that the modes in the stub are in this case the coaxial equivalents ($TE'_{1,1}$ and TEM) of the $TE_{1,1}$ and $TM_{0,1}$ modes. (See figure 26.) A resonant ring is used in case the stub is not completely efficient. The ring's position

is not important as long as it does not form a resonant section with the stub. Figure 52 (C) perhaps needs some explanation. It is a British scheme for obtaining the $TM_{0,1}$ in round guide by transforming

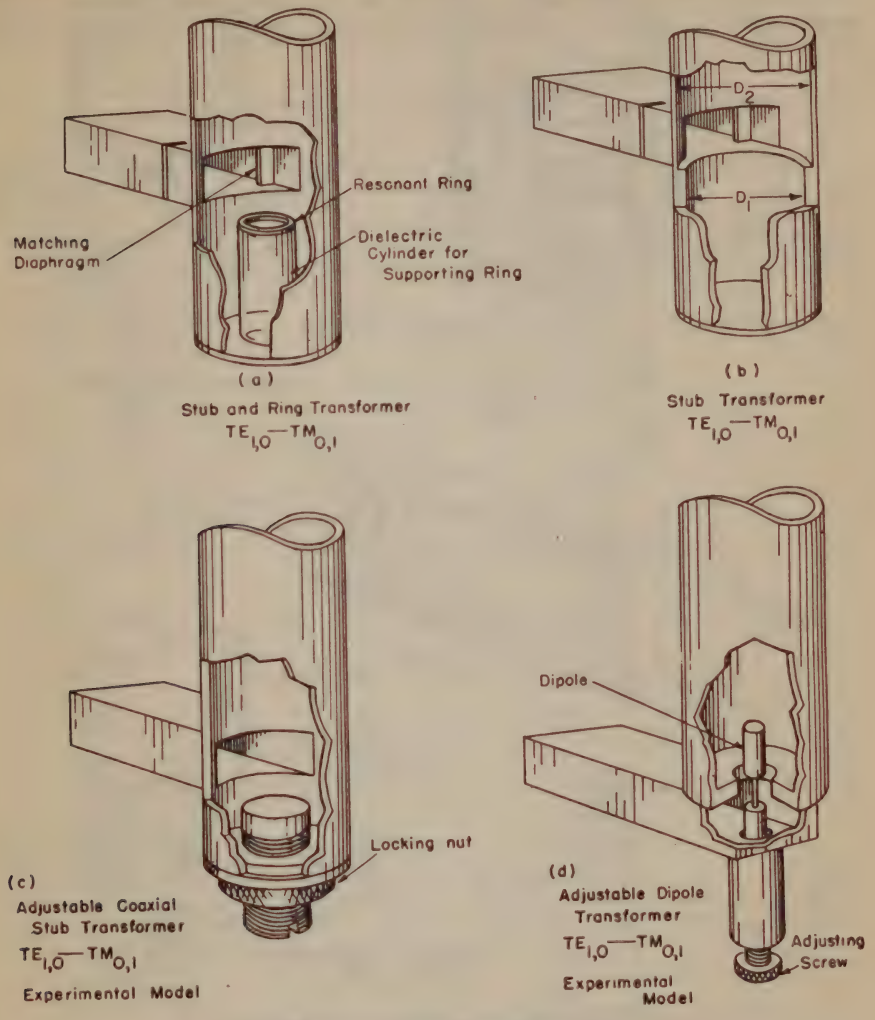


FIGURE 51.—Rotary transformers.

from $TE_{1,1}$ in round guide. In the larger guide half of the $TE_{1,1}$ mode goes into the dielectric material and half does not. Since the wave travels slower through the dielectric than through the air, it is possible to choose the length so that at the end of the dielectric the two halves are 180° out of phase. The resultant field then approximates the $TM_{0,1}$ mode. The edges of the dielectric are curved to avoid corona.

The second kind of coupling is dipole coupling and is illustrated by figure 51 (D).

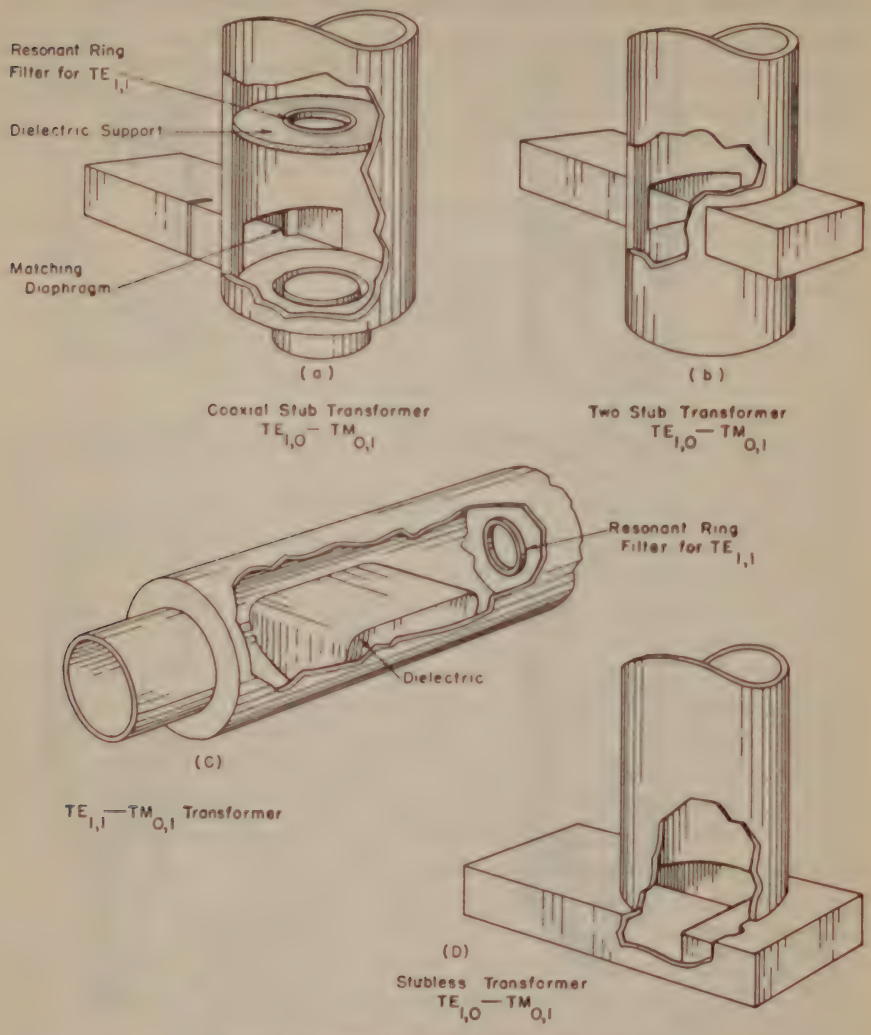


FIGURE 52.—Rotary transformers (additional types).

i. Coaxial Line to Wave Guide Transformations.

A simple and broad band-pass transition is of the type shown in figure 53 (A). The center conductor extends into the wave guide approximately a quarter wavelength. Note the similarity to a current-fed quarter-wave vertical antenna of broadcast stations. The end plate on the wave guide is approximately a quarter of a guide wavelength from the antenna, so that it reflects radiation in phase with that radiated directly down the guide.

In actual practice neither of these lengths is a quarter wavelength, but must be determined by experiment. For example, the dimensions for the 1⅝-inch coaxial line, 50 ohms, to 1½-inch x 3-inch wave-guide transformation best matched at 10.2 cm. ($r=1.05$) were as follows: antenna length, 0.18λ air and end-plate distance 0.16λ guide. Such

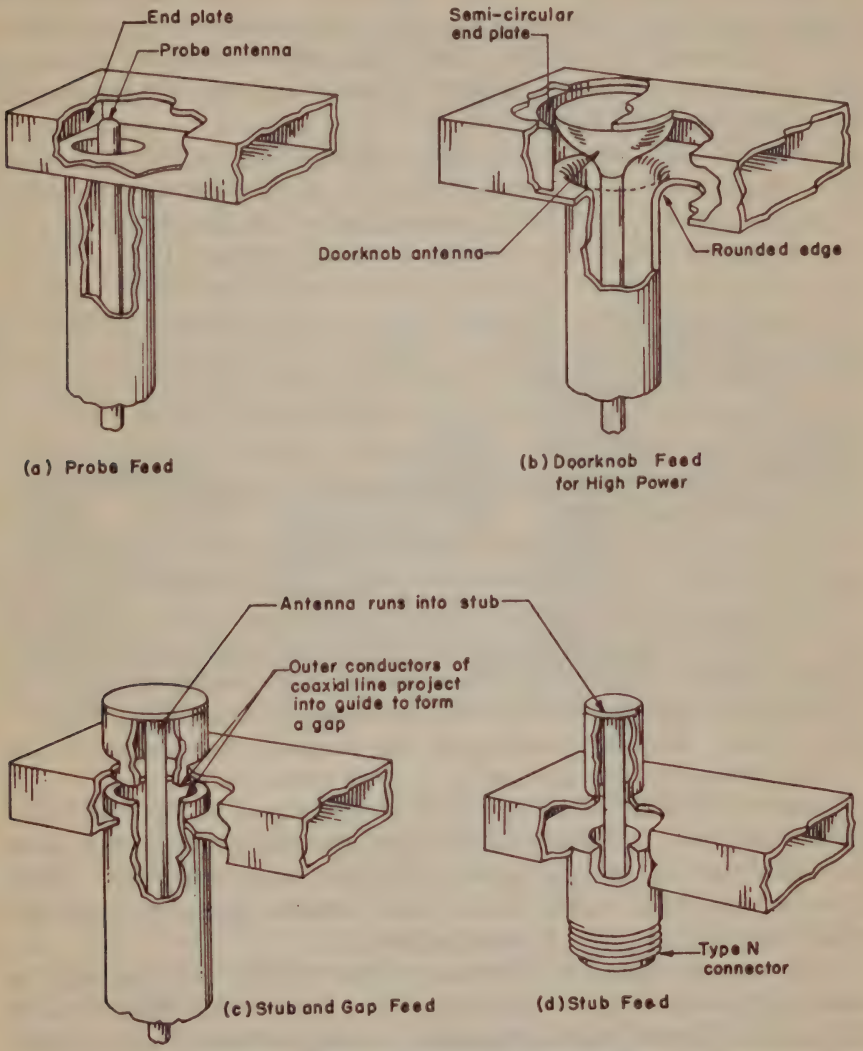


FIGURE 53.—Coaxial line to wave guide transformations.

a device had $r<1.1$ from 9.7 to 10.8 cm. and $r<1.2$ from 9.2 to 11.5 cm.

The reason for the departure from quarter wavelength spacing is partly due to the usual difference between measured and effective lengths which occurs when line size becomes comparable to a wavelength.

The discussion of Slater ("Microwave Transmission," section 39) suggests a more fundamental reason. We may consider the end-plate distance as varying the amount of coupling between coaxial and guide (through phase of directly radiated and end-plate-reflected fields). This enables one to match the characteristic impedance of the guide to that of the coaxial except for the reactive component of the transition. This reactance may be matched out by adjustment of the antenna length. Naturally neither of these variations gives either pure resistance or pure reactance changes, but the general viewpoint may have some merit.

Figure 53 (D) shows a variation, which may be thought of as being the feed shown in figure 53 (A) with the quarter-wave antenna extending through and being stub-supported in the opposite wall. Or, perhaps better, it may be considered as forming a large coupling loop linking all the flux between the coaxial center conductor and the wave guide end-plate. Varying the end-plate distance varies this linkage, and hence the degree of loading of the coaxial line by the wave guide (or vice versa). By varying the length of the stub support the reactive component of the loading may be tuned out.

Unfortunately this type of transition is several times more frequency sensitive than one of the type shown in figure 53 (A) so is not good for broadband work.

Figure 53 (B) shows a feed which is more complicated analytically than the others. In performance it seems quite satisfactory. It is roughly the same as figure 53 (A) as regards frequency sensitivity while at the same time it has greater ruggedness and probably greater power handling capabilities. Analytically it is similar to figure 53 (D) with the shorting plug of the stub support protruding into the wave guide. The streamlining of the contours reduces the fields, making this one of the best units for high power work.

Another modification of figure 53 (D) is shown in figure 53 (C). There the outer conductors are brought into the guide, giving as an additional variable the gap between the two outer conductors. Such a transition is used in the 3 cm. band tunable probe, the gap and coaxial shorting stub both being adjustable.

As mentioned above in discussing figure 53 (D) coupling may be made by terminating the coaxial line in a loop which links some of the magnetic flux of the wave guide. A simple loop of more conventional type might be used, but in order to obtain an impedance match

it would have to be very large. This would make manufacture and assembly difficult. The transformation would probably be much more frequency sensitive than the other types discussed, and would be likely to break-down under high power.

j. Rotary Joints.

It is evident that in order to avoid polarization effects, a rotating joint requires axial symmetry in the electromagnetic field. This requirement is the only reason at present for ever using a mode other than the lowest in wave guide. The $TM_{0,1}$ in circular guide is the lowest mode with axial symmetry and is for that reason used in rotary joints. Two $TE_{0,1}$—$TM_{0,1}$ transformers (such as described in section II, 3h) connected by a circular guide whose two sections rotate with respect to one another comprise a rotary joint. The only limi-

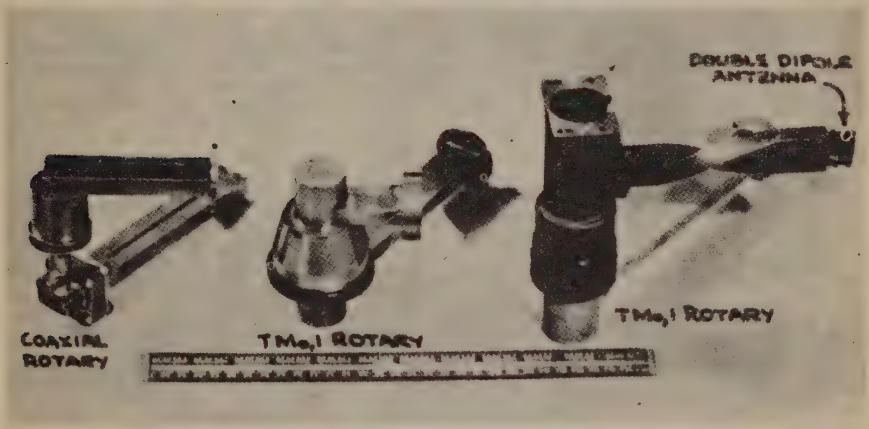


FIGURE 54.—3 cm. rotary joints.

tation on this joint is that the distance between the rectangular guides must not be a resonant length for the $TE_{1,1}$ mode at any wavelength within a certain band. This limitation causes certain regions of lengths which are not permitted for rotary joints which are good over a 2 percent band. As the distance between the rectangular guides increases, these regions spread out, and when the distance is about 10 guide wave lengths long the resonance regions begin to overlap at the edges of the 2 percent band. Thus a distance of about 40 cm. is the maximum length for a wave guide rotary joint in the 3 cm. region which is to perform satisfactorily over a 2 percent range. The corresponding distance at 10 cm. is about 120 cm. The two guides are capacity coupled in just the same manner as the outer conductors in the coaxial rotary joint described in section II, 1e. (See the $TM_{0,1}$ rotary joints in figure 54.)

Another type of rotary joint utilizes a short section of coaxial line

connecting two rectangular guides. 3 cm. rotary joints of this type are shown in figure 54 (identified as "coaxial rotary"), figure 55, and figure 56. The rotating part differs from the usual coaxial rotary joint in that the central conductor is all one piece.

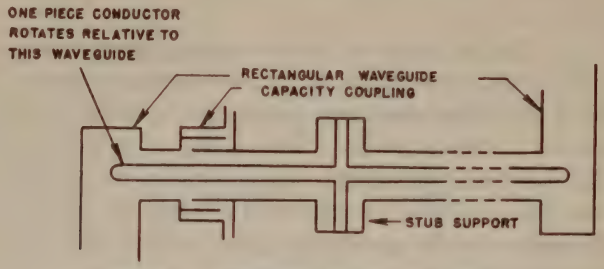


FIGURE 55.—Coaxial rotary joint for wave guide.

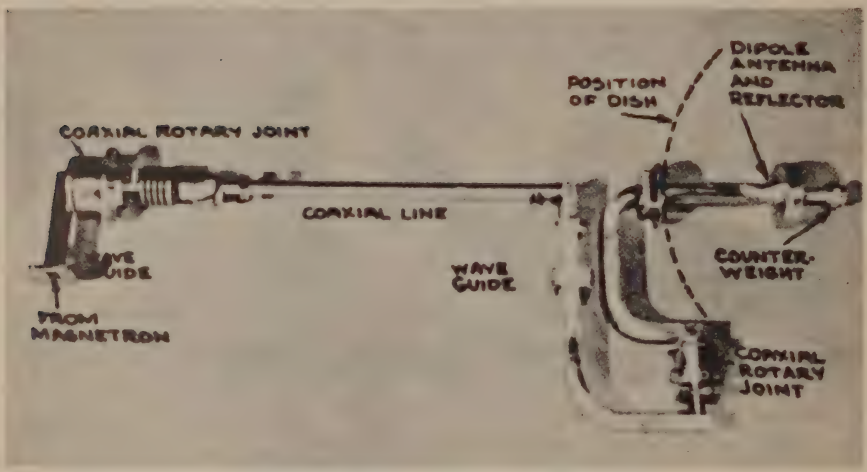


FIGURE 56.—Sample 3 cm. spinner plumbing.

III. R-F COMPONENTS OF 10 CM. AND 3 CM. RADAR SYSTEMS

1. MAGNETRON COUPLING

The power output and operating frequency of 3 and 10 cm. magnetrons follow a fairly definite pattern with respect to the load impedance regardless of the input voltage and current. Consequently, the loading may be considered without reference to the input conditions. The frequency itself is not a critical variable since provision is made for tuning the receiver. However, it is important that the frequency be stable, that is, it should not be "pulled" too greatly by the accidental variations in line termination which are unavoidable in an r–f system containing moving parts.

Generally the output impedance which leads to maximum efficiency results in an objectionable degree of frequency pulling, so that a compromise should be made in the load. The proper choice is apparent only after a systematic and rather long experiment has been made. 10 cm. production magnetrons are sufficiently uniform so that for a given type the match may be made on the basis of laboratory data from a few sample tubes. 3 cm. tubes are at present supplied with a coupler adjusted at the factory.

It is customary to express the degree of frequency pulling as the "pulling figure." This is defined as the maximum change in frequency in megacycles which occurs when the magnetron is made to operate into a load of $r=1.5$ and the position of the minimum is varied through a half wavelength. This load condition is sometimes expressed as an r of 1.5 of all possible phases. Both the pulling figure and efficiency depend on the load impedance. Average values of pulling figures and efficiencies for S-band magnetrons at an operating point of approximately 10.5 kv. and 12.5 amperes are:

Tube	Pulling figure	Efficiency percent
2J32	4	35 to 40
2J25	8	29 to 33
2J22	14	30 to 35.

At an operating point of approximately 20 kv. and 30 amperes, the efficiencies for all of these tubes is about 45 to 50 percent. For 3 cm. tubes, the pulling figure is about 9 and the efficiency is about 12 percent for the 2J21 and 40 percent for the 725A.

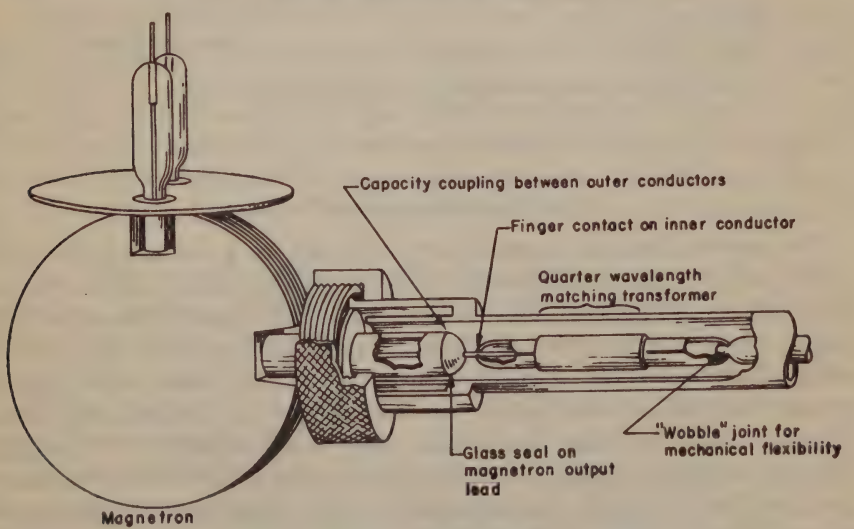

Capacity coupling between outer conductors
Finger contact on inner conductor
Quarter wavelength matching transformer
"Wobble" joint for mechanical flexibility
Glass seal on magnetron output lead
Magnetron

FIGURE 57.—Typical magnetron coupling to coaxial line.

The fact that so far the only satisfactory way of loading the magnetron cavity is by means of a coupling loop has the consequence that the output lead is in the form of a coaxial line, which must be connected to a coaxial transmission line, or to a coaxial-to-wave-guide coupler. In either case, the problem is one of joining coaxial lines of different dimensions and adding the appropriate quarter-wave transformer to the center conductor to secure the chosen load impedance. The principal difficulty encountered in the design is that of avoiding power limitation due to arcing. A typical coupler is shown in figure 57.

2. THE TR BOX

a. Purpose and Theory of the TR Box.

A TR box or a transmit-receive switch is an electronic switch which enables a radar system to use the same antenna for both transmitting and receiving. A switch is necessary to prevent the transmitted power from injuring the sensitive receiver, on one hand, and to prevent the received signal from being absorbed in the transmitter, on the other.

Fortunately the second purpose is easy to achieve since a magnetron usually changes its impedance a great deal between the transmitting and quiescent state. To protect the receiver is a much more difficult matter. It is necessary that the switching device act within a fraction of a microsecond and that it provide an excellent short circuit for the receiver: the peak power into the receiver must be of the order of 60 db below the transmitted power. Some form of gas discharge is generally used. The simplest type might take the form shown in figure 58 for the case of a parallel line system.

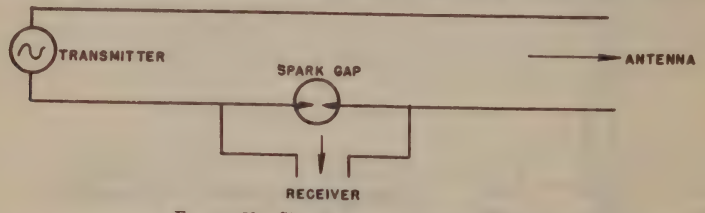


FIGURE 58.—Simple transmit-receive switch.

The power from the transmitter will build up a voltage across the gap until it breaks down; most of the power will go out to the antenna, and only a small fraction to the receiver. The signal power, on the other hand, will not be sufficient to cause a discharge and, if the transmitter has a low impedance when quiescent, most of the signal power will go into the receiver. The power to the receiver when the gap is fired is called the *leakage power* and is equal to V^2/Z, where V is the voltage across the gap and Z the impedance of the receiver measured at the gap. The space around the gap may be evacuated to a reduced pressure so that the voltage V is lowered.

This simple scheme could be employed in either coaxial line or wave guide, but it does not offer sufficient protection to the receiver. *The obvious improvement is to insert some voltage step-up transformer before the gap and a step-down transformer after it,* and this is, in effect, what is done. In the microwave region, the analogue of such an arrangement is a resonant cavity in which the gap is placed. Such a cavity may take many forms and may resonate in many modes but the simplest and most commonly employed shape and mode is the following. Suppose we have a cylindrical box with perfectly conducting walls with two posts along the axis of the cylinder forming the gap, as shown in figure 59. In this case, the lowest mode of oscillation is the one of interest to us. In this mode the electric field is along the axis of the cylinder and increases as we go from the edge of the box to the center. The magnetic field lines are circles perpendicular to the axis of the cylinder, and the currents flow radially and down the center posts.

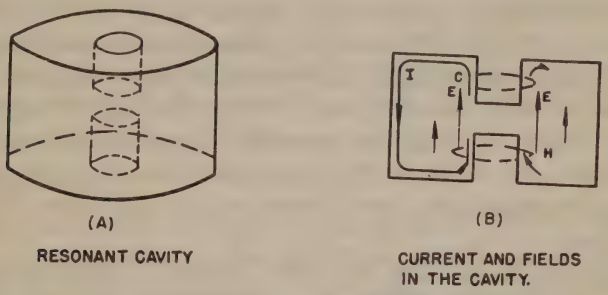


FIGURE 59.—Resonant cavity TR box.

We may couple energy into and out of such a cavity either by holes in the cylindrical walls or by inserting loops into the cavity as shown in figure 60. These two methods of coupling are entirely equivalent and may be represented by the equivalent circuit of figure 61. L' is the small inductance of the short length of center post and C' is the capacity between the posts. R represents the I^2R losses in the cavity walls and posts. The transformers represent the coupling in and out.

The step-up ratio of the transformer is related to the size of the coupling windows or loops, the ratio being higher the smaller the windows or loops. An alternative method of representation would be to employ 1 : 1 transformers whose mutual inductances between windings depend on the degree of coupling to the cavity. The two methods are entirely equivalent.

The resonant cavity can be characterized by a value of Q in exactly the same way as the equivalent resonant circuit. We have $Q = R_{eff}/X_{eff}$, where R_{eff} is the effective shunt resistance of the circuit, and X_{eff} the

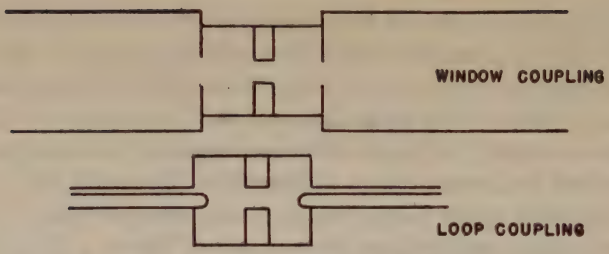


FIGURE 60.—Coupling in a TR box.

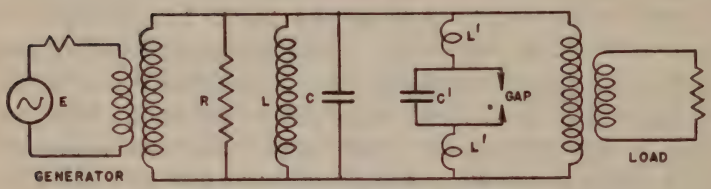


FIGURE 61.—Equivalent circuit of a TR box.

inductive reactance at resonance. R_{eff} is made up of three parts: the contribution from the resistance R shown in the circuit above, and the contribution from the load and generator impedances transformed through the step-up transformers. If the coupling through the TR box is very small, the load and generator impedances have no effect, and Q in that case, which is R/X_{eff}, is called the unloaded Q. It is a useful quantity since it is a good measure of R and hence of the losses in the TR box.

When the gap is not fired, and the TR box is tuned to resonance, power will be transmitted from the generator to the load with some loss in the resistance R. The equivalent circuit reduces to figure 62 (A).

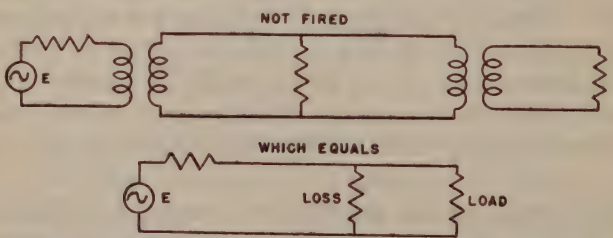


FIGURE 62 (A).—Equivalent circuits of a TR box when not fired.

When a discharge takes place across the gap, the voltage across the gap is limited and the power to the load is diminished greatly. This leakage power is composed of two parts. The first portion is proportional to the square of the voltage across the gap and this is called the normal mode coupling through the TR box. The second part is independent of this voltage. It is proportional to L' and the generator

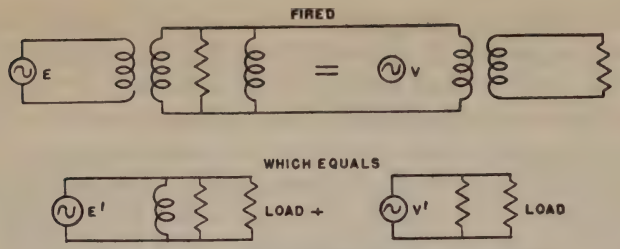


FIGURE 62 (B).—Equivalent circuits of a TR box when fired.

power, and is called the direct coupling. At low generator powers the normal coupling is important, but at high powers the direct coupling becomes the principal effect. The equivalent circuit may be reduced to those shown in figure 62 (B). The output power naturally depends upon the impedance of the load and may be shown to be inversely proportional to the real component of this impedance. The leakage power also depends upon the loaded Q of the TR box. The lower the Q_L the lower the value of E_{eff}, since the load and generator impedances make a larger contribution, and the higher the leakage power. Using the equivalent circuits shown in figure 62 it is possible to calculate accurately the exact form of all the interrelations between the various quantities involved.

b. 10 cm. TR Boxes.

The TR box allows one to use a common antenna for transmitting and receiving. However, the saving in antenna space and weight must not be accompanied by a large loss of received signal. If the TR cavity is properly adjusted the losses through it may be kept to an arbitrarily low value by a choice of the input and output loops. The losses in the transmitter may also be kept low if certain precautions are taken.

Two TR tubes are used for the 10 cm. band, the 1B27 tube which is tuned by changing the gap spacing, and the 721A tube which must be used in a cavity provided with tuning plugs.

Depending on the method of coupling the TR to the line, it is either in series or in shunt with it (figure 63). If the shunt connection is

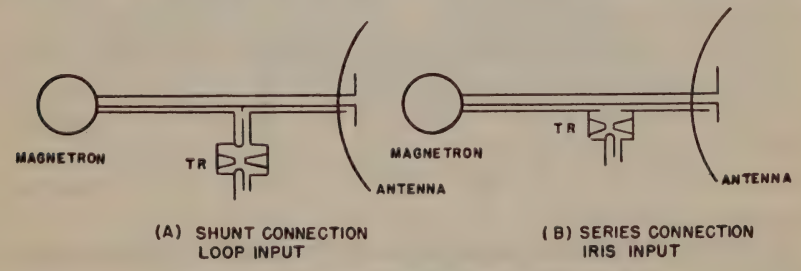


FIGURE 63.—TR boxes in shunt and series with the line.

used, the *admittance* at the junction due to the magnetron must be low, so that the larger part of the received signal goes to the TR box and on into the receiver rather than being absorbed in the magnetron cavity. If the series connection is used, the *impedance* at the junction due to the magnetron must be low for the same reason.

It is a happy circumstance that most 10 cm. magnetrons have greatly different r-f impedances under oscillating (hot) than under quiescent (cold) conditions. As a result, it is possible to match the magnetron to the line so that maximum power is extracted when it is oscillating and also so that it presents a bad mismatch to a received signal when it is quiescent. Thus, on the received signal, large standing waves are set up, and the TR box may be set at a maximum (low admittance) or a minimum (high admittance) of the voltage standing wave, depending upon whether shunt or series coupling is used. When the junction point is properly chosen, the loss in the magnetron may be held down to less than 0.5 db. Tests have shown it is possible to choose a fixed distance that is suitable for all S_A magnetrons, another for all S_G and another for all S_S magnetrons.

There are installations where it is not convenient to place the TR box as close to the magnetron as the above methods require; or it

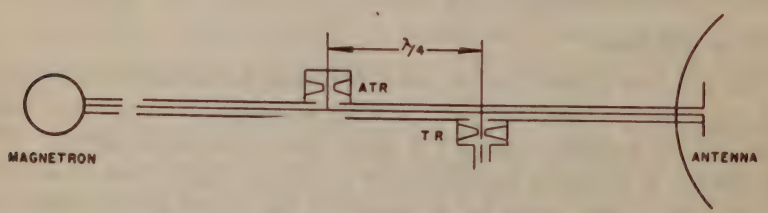


FIGURE 64.—ATR and TR boxes in series connection on a line.

may be necessary to use magnetrons with different cold impedances. In any event, it is possible to place a TR box at an arbitrary point on the main line if an anti-TR (ATR) is used in conjunction with it. The ATR is simply a standard TR tube in a cavity with no output circuit which is placed on the line between the TR and the magnetron. When the magnetron transmits, its gap breaks down (as does the TR), and the transmitter power goes right by the ATR. For received (low power) signals, the ATR does not fire, and either shorts the line (if it is in shunt) or open circuits the line (if it is in series) and sets up a large standing wave. The TR is then placed at the appropriate point in reference to the ATR. Such a combination may be placed at any point along the line without reference to the magnetron impedance (figure 64).

Figures 65 and 66 are photographs of a number of different methods of coupling a TR cavity to a line.

Crystal protection.—In the transmitting state the function of the TR tube is twofold. One, is to make "all" of the magnetron power go down the line to the antenna; the other, is to keep the line power from getting into the receiver and burning out the first detector

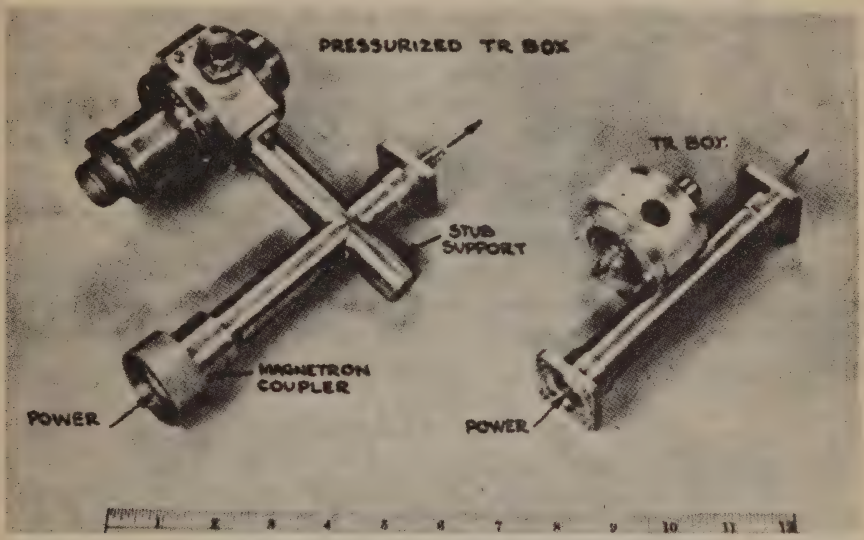


FIGURE 65.—Ways of coupling 10 cm. TR boxes to aline. Left, coaxial T-loop coupled in shunt with the line. Right, coaxial iris coupled in series with the line.

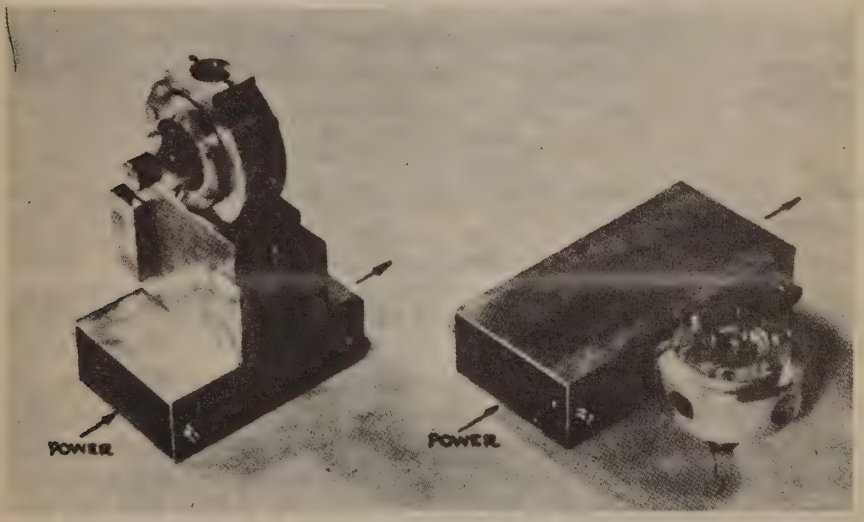


FIGURE 66.—Ways of coupling 10 cm. TR boxes to a line. Left, wave guide T-iris coupled in series with the line. Right, wave guide iris coupled on the narrow side in shunt with the line.

(crystal or vacuum tube). The 721A tube is designed to protect crystals, and as such must limit the peak leakage power to considerably less than one watt. A desirable figure is something less than 150 milliwatts.

The envelope of the r-f energy coming through a TR tube is shown in figure 67. Since the operation of the TR tube depends on the electrical break-down of the gap between two electrodes, a certain "over-voltage" must be applied to the gap in order to fire it in a very short time (less than 0.01 microseconds). It is this over-voltage that produces the "spike." The "flat" represents the leakage power due to the arc voltage coupling to the output loop (arc coupling) and the exponentially attenuated higher modes directly from the input to the output loops (direct coupling).

FIGURE 67.—Envelope of pulse through a TR box.

Experimental data indicate that for a 721A tube in a cavity whose input and output loops are adjusted for a transmission loss of 1.5 db whose unloaded Q is 2000 and which presents a matched load to the main line, the leakage power in the flat is expressed by:

$$P_F = P_{ARC} + P_{D.C.} = 0.04 + \frac{P_L}{4 \times 10^6}$$

where P_F = leakage power in flat,
P_L = transmitter power.

The arc coupling is essentially constant for all line powers from 1 kw. to 500 kw. peak; and the direct coupling is directly proportional to, and 66 db down from, the line power.

The effect of changing the transmission loss through the cavity (keeping the TR a matched load to the line) is shown by the following formulas:

$$P_{ARC} \sim \frac{T}{(1-T)Q_0}$$

where T = transmission through the TR (output/input), and Q_0 = unloaded Q of the cavity.

$$P_{D.C.} \sim \frac{1}{Q_0^2} \frac{T}{(1-T)^2}$$

Figure 68 is a plot of the variation in arc coupling and direct coupling as a function of transmission loss.

The leakage power in the spike has not yet been accurately measured. Its duration is so short (less than 0.1μ sec., and probably less than 0.02μ sec.) that the video amplifiers in r-f envelope viewers cannot respond to it. Various estimates have been made, and it is generally agreed that the spike may be at least 200–400 mw. greater than the flat.

Turn on.—The initiation of a gas discharge between two electrodes requires the presence of a few ions in the gap at the time voltage is applied. If there are none present, the voltage must rise to a value great enough to cause "cold emission" from the metal. Because of this, there is a large probability that the first few pulses may not break

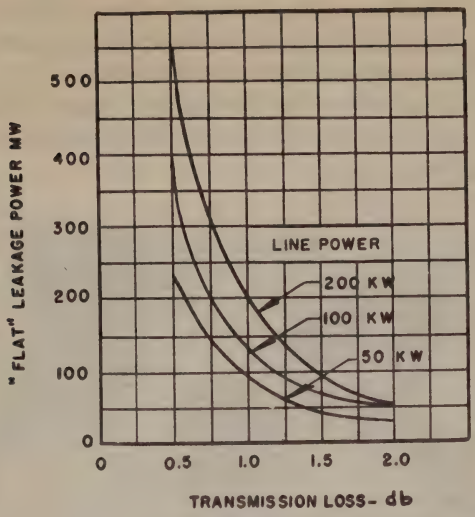


FIGURE 68.—Leakage power through a 721-A TR box as a function of transmission loss (matched input, $Q_0=2000$).

down the gap in the TR tube when the system is turned on. If the tube does not fire, the leakage power may be half the line power, and will certainly burn out the crystal. To forestall such an occurrence, an auxiliary "keep alive" electrode is built into the tube, figure 69. A negative d–c potential is applied to this electrode such that a discharge is started (the current is limited to 0.1 to 0.2 ma.). This is turned on before the main r–f power, and thus furnishes ions for the first few pulses, allowing the main gap to break down immediately at a normal voltage. The keep-alive also reduces the spike if a negative voltage is used.

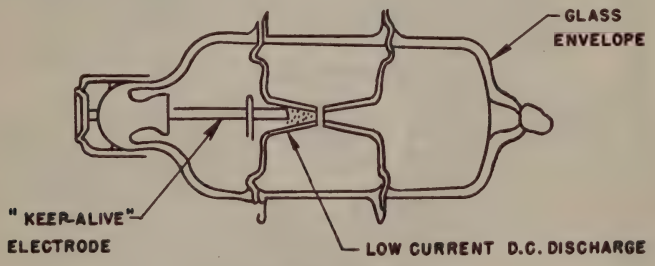


FIGURE 69.—TR tube.

Since the time between application of the keep-alive voltage and the turning on of the r–f power may be as short as 20 sec., it is important that the keep-alive start in within a few seconds. The momentary application voltage much higher than normal would make the keep-alive time lag short. However, such high voltages are not always convenient to obtain. A simpler way is to furnish a primary

source of ions. This takes the form of some radioactive salt (radioactive cobalt), either applied to the back of the cone as a paste, or mounted in a disc on the keep-alive electrode.

Typical cavities are shown in the figures 70, 71, 72 and 73. Their tuning ranges vary from 10 to 15 percent depending on the size and number of tuning plugs used.

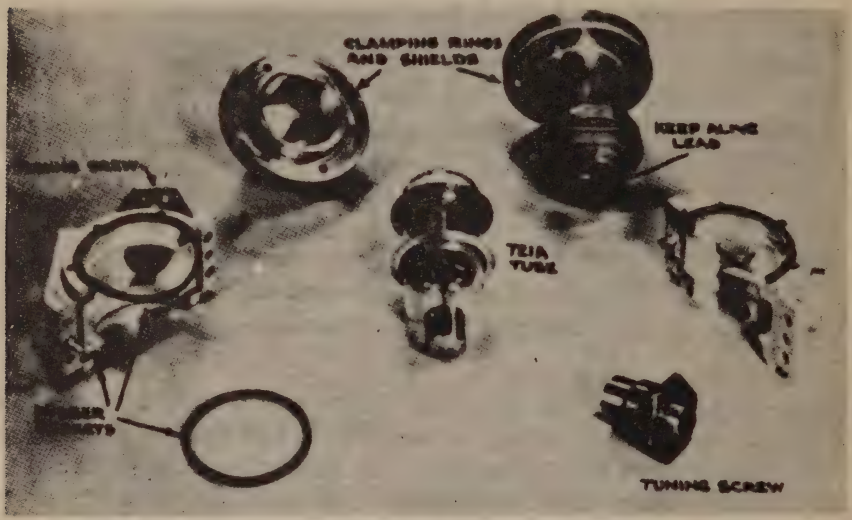


FIGURE 70.—Exploded view of a pressurized TR box.

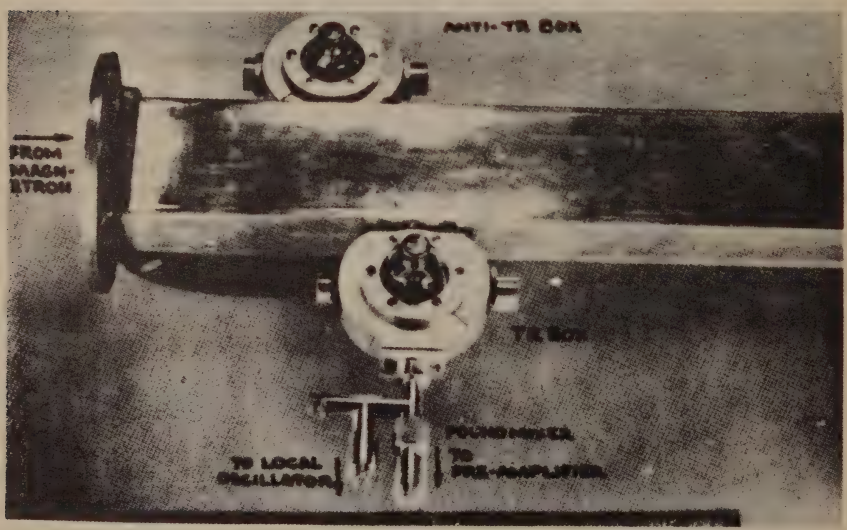


FIGURE 71.—10 cm. wave-guide duplex-T assembly.

c. 3 cm. TR's.

Introduction.—3 cm. TR boxes are of two types: the 1B24 (integral unit) and the 724A or 724B tube with separate cavity.

The integral unit type consists of a cylindrical cavity with two conical center posts almost meeting in the center (see figure 74).

An experimental model of the 1B24 is shown in figure 73. This cavity is coupled to the guide by circular windows. Glass is sealed across the windows, so the entire cavity may be filled with a low-pressure gas mixture. Tuning is accomplished by varying the gap

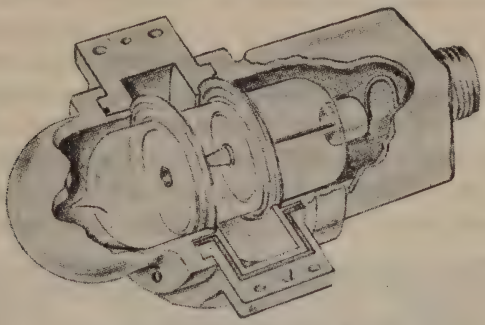


FIGURE 72.—Pressurized 10 cm. TR box.

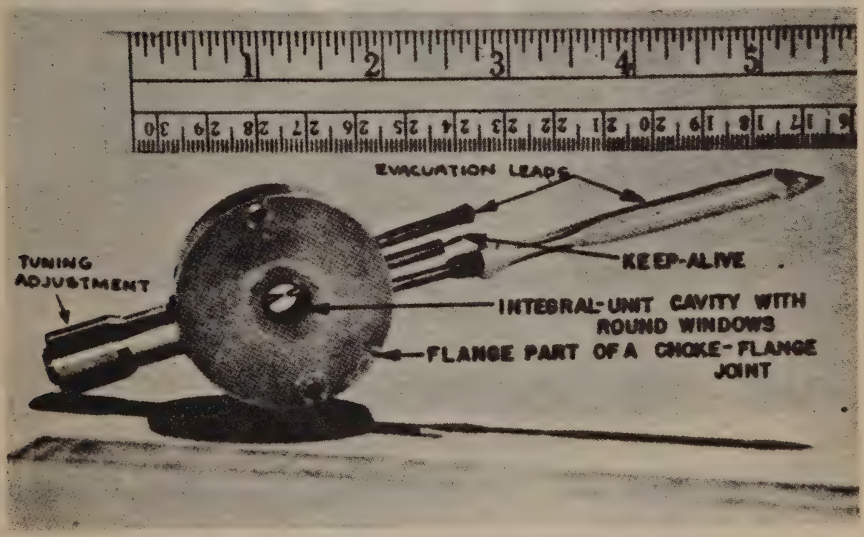


FIGURE 73.—Westinghouse 3 cm. TR box (experimental model).

between the center posts. A third electrode, insulated from the cavity, is inserted inside the hollow lower electrode. A d-c potential is applied between this electrode, the "keep-alive," and the cavity, with the keep-alive negative. This sustains a continuous discharge between the keep-alive and the inside of the cone. Some of the ions from this discharge wander into the space between the center posts, and cause the gap to break down more readily when an r-f pulse comes along.

FIGURE 74.—Integral Unit—TR cavity.

The other type of TR box is in two parts: A gas-filled glass tube (type 724A) containing the conical electrodes and the keep-alive and a split metal block forming the rest of the cavity, the coupling windows, and the tuning screws. A picture of the cavity is shown in figure 75.

The cavity of figure 75 is tuned by moving the plungers in the side. One tuner is a screw covering a wide tuning range. This cannot be used for continuous tuning, but is set near the right place and locked. When used thus it introduces little loss into the cavity. The fine tuner, which does not touch the sides of the cavity, is used to get the

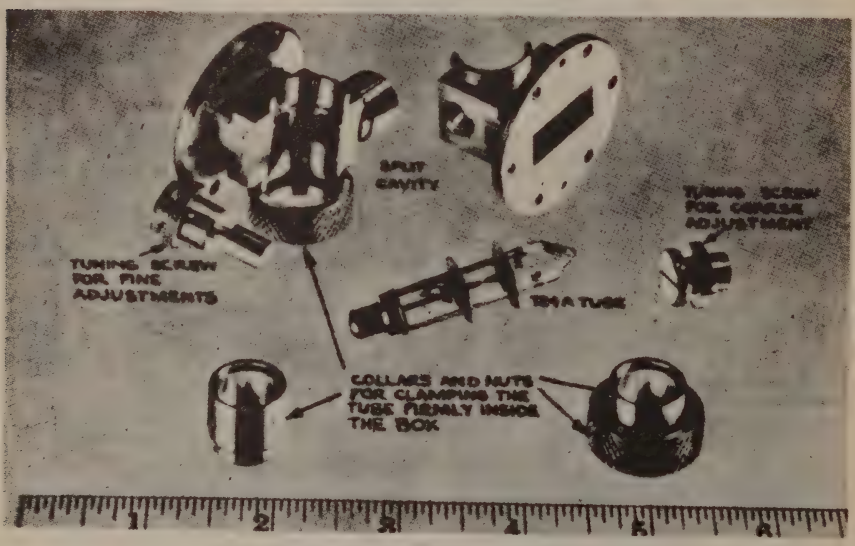

SPLIT CAVITY
TUNING SCREW FOR COARSE ADJUSTMENT
TUNING SCREW FOR FINE ADJUSTMENTS
COLLARS AND NUTS FOR CLAMPING THE TUBE FIRMLY INSIDE THE BOX

FIGURE 75.—3 cm. TR box and tube

exact frequency desired. Rectangular coupling windows are used in this type of TR box.

Properties of the TR box.—(*1*) *Tuning range.*—Two cavities designed for the 724B tube tune from 3.13 to 3.28 cm. and from 3.22 to 3.39 cm., respectively. The 1B24 TR tunes from 3.1 to 3.5 cm.

(*2*) *Q and loss.*—The intrinsic, or unloaded Q, Q_0, of the Western Electric type TR is around 1250; that of the 1B24 is around 2400. The Q_0 of the former TR is lower partly because the lossy glass is inside the cavity where the field is higher. The 1B24 has the glass at the edge of the cavity. The W. E. type TR has a loaded Q, Q_L, of 200 with a cavity loss of 1.8 db. Less cavity loss would be desirable, but this can only be done at the expense of greater high-level leakage

power, which is hard on the crystal. The 1B24 has a cavity loss of 1.1 db., and a Q_L around 330.

(*3*) *High-level leakage.*—The W. E. type TR passes from 20 to 60 milliwatts peak power when the magnetron is supplying 10 kw. peaks. The 1B24 will pass about 8 mw.

(*4*) *Recovery time.*—A certain time is required after an r-f pulse has fired the gap in the TR cavity for the ions to recombine. While the ions are still around in sizable numbers they will absorb an appreciable fraction of the energy of a signal passing through the TR cavity so that signals coming shortly after the TR has stopped firing will be reduced in intensity. In addition, a cloud of ions in the cavity will change the dielectric constant of the cavity, thus detuning it. How much loss is contributed by each of these effects is hard to say. We define the recovery time as the time after the end of the gap discharge when a received signal is 3 db. down from one received a long time after the ions have recombined. A unused TR has a recovery time of 1 to 2 microseconds.

(*5*) *Life.*—After a TR has been in operation for some time, changes take place in its properties which make it useless. The principal change is a reduction of the gas pressure by absorption into the tube wall, in the case of the W. E. type, TR, or the cavity wall in the case of the 1B24. This results in an increase in the recovery time and, after a longer time, an increase in the high-level leakage. There may also be some sputtering of the electrodes, resulting in the deposition of a metallic film on the glass surfaces in the cavity, and a resultant increase in the loss in the TR, and a lowering of the Q. The gas clean-up is apparently caused mainly by the keep-alive discharge. This means that the keep-alive current should be kept as small as possible consistent with adequate crystal protection. Lowering of the cavity Q because of sputtering has occurred only in one known case, and this was after long operation.

(*6*) *Various troubles.*—In some cases TR boxes allow the first few of a string of high-level r-f pulses to pass through the cavity without firing the gap. This is known as the turn-on effect, and is hard on crystals. For this reason TR boxes are provided with a "gate" operated by a solenoid so that the crystal is disconnected from the TR until after turnon.

(*7*) *Spikes and the importance of keep-alives.*—It is now believed that the preignition energy (spikes, see figure 67), which is transmitted through the TR box before the tube fires, is responsible for most crystal failures. Experiments on the 724A tube indicate that approximately one-third of the total leakage power at 10 kw. operation is contained in the spike. Since the duration of the spike is extremely short (of the order of 10^{-9} sec.), the peak spiked power will be much greater than the flat peak power.

X band TR boxes may pass spike energies of the order of 5×10^{-9} joules per pulse. This is with keep-alive completely effective; this precaution insures that all spikes have the same aplitude. The 1N21A crystal is required to withstand spike energies of 3×10^{-8} joules at manufacture.

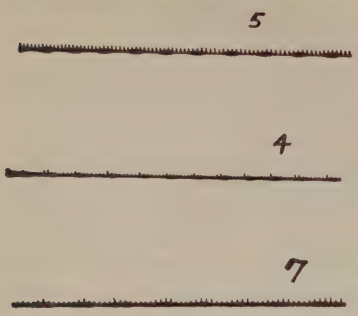


FIGURE 76.—Top, successive spikes when keep-alive is functioning properly. Center, successive spikes when keep-alive is oscillating. Bottom, successive spikes as the keep-alive is turned off (photographed at 500-cycle repetition rate).

(*a*) *Keep-alive effectiveness.*—As explained previously it is necessary to have some ions in the gap when the r-f voltage is applied to insure rapid formation of the arc. The keep-alive electrode is built into the tube to supply these ions. When the keep-alive is operating properly so that the required ions are always present successive spikes will be uniform and small as shown in the top line of figure 76.

If the keep-alive is off or not completely effective the total spike energy will be greater and the spikes will have a random energy distribution, some of them being exceedingly large (bottom line of figure 76). The center line of figure 76 shows the spike amplitudes when the keep-alive is oscillating so that it is off part of the time.

On systems operating at high-recurrence rates there is sufficient carry-over of ionization to insure rapid formation of the arc, but at low repitition rates this is not so. Figure 77 shows how the average spike energy varies as a function of repetition rate when the keep-alive is not connected.

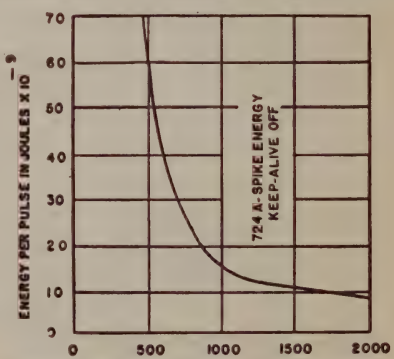


FIGURE 77.—Spike energy as a function of repetition rate, in cycles per second.

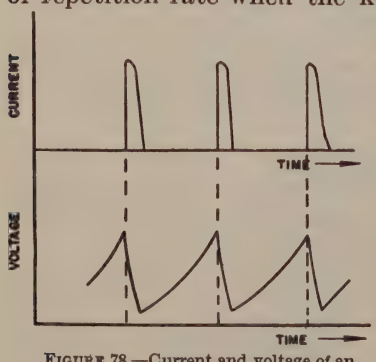


FIGURE 78.—Current and voltage of an oscillating keep-alive.

(*b*) *Oscillating keep-alive.*—Over certain sections of its volt-ampere characteristic the keep-alive electrode operates as a relaxation oscillator. The current pulse of such an oscillator is shown in figure 78. The minimum current where keep-alive oscillations are not troublesome is a function of the external supply circuit, the keep-alive electrode design, and the gas filling.

(*c*) *Circuit design.*—The equivalent circuit for the keep-alive is shown in figure 79, where R is the current limiting resistor and C is the capacity of the electrode plus the distributed capacity of the lead from the limiting resistor to the tube. For satisfactory operation, C should be made extremely small. This can be accomplished by mounting the current limiting resistor R at the electrode terminal as shown in figure 80. Since oscillation is also a function of the keep-

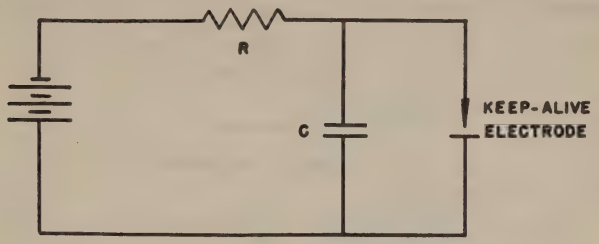


FIGURE 79.—Equivalent circuit for the keep-alive.

alive current, occuring at low values of current, it is necessary to use a well-regulated power source to supply the keep-alive. Care should also be taken to see that switching transients and arcs within the modulator do not cause the keep-alive voltage to drop momentarily.

Anti-TRs and duplexing assemblies.—We must consider now how the TR is used in a system. Figure 81 shows in simplified form the essential features of the usual arrangement.

When a high-level pulse comes down the line from the magnetron some of the energy enters the TR cavity, causing the gap to fire. The arc detunes the cavity. Looking into a cavity off resonance we see a short across the line at the window. If the TR window is the proper distance from the main line this short will appear across the entrance to the T arm, so essentially all of the magnetron power will pass by the TR line to the antenna. The received signal from the antenna does not fire the TR, so it sees a matched line looking at the TR, and some of the power goes up this line, and into the receiver. However, some of the signal will be lost into the magnetron, the amount that is lost being dependent on the impedance of the magnetron when cold and the length of the line leading to the magnetron. We can prevent this loss by putting in an anti-TR box between the TR line and the magnetron, as will be shown.

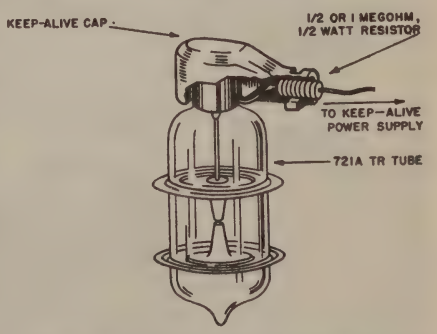


FIGURE 80.—Mounting of the limiting resistor in the keep-alive circuit.

The anti-TR is like a TR with only one window. If the anti-TR is placed in a side arm, as the TR is, a high level pulse from the magnetron will fire the tube. If the window is the proper distance from the main line the power will pass by the side arm as it does in the case of the TR. However, a signal coming in from the antenna will not fire the tube and will be reflected from the anti-TR junction.

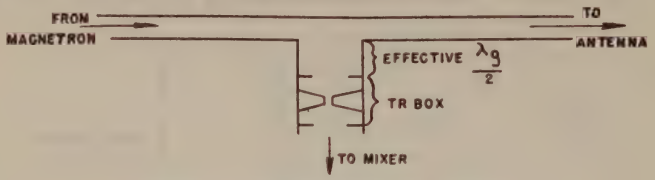


FIGURE 81.—TR assembly using an *E* type T.

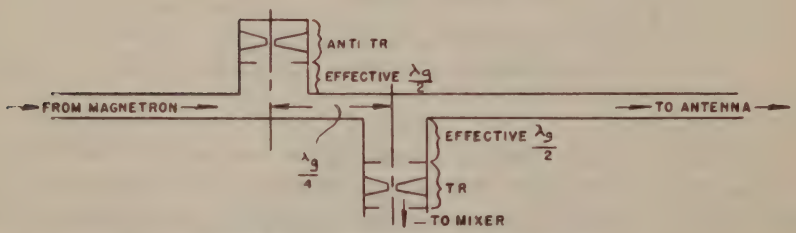


FIGURE 82.—Duplexing assembly using *E* type T's.

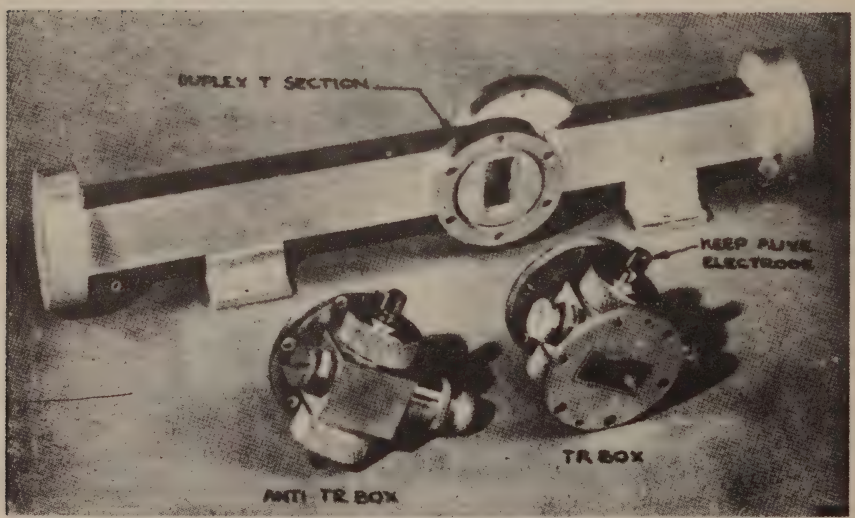


FIGURE 83.—3 cm. duplex I assembly.

Now if the distance from the anti-TR to the TR be correctly chosen, a maximum signal power will pass around the TR junction and through the TR.*

The complete assembly which is shown in figures 82 and 83 is known as the duplexing assembly.

*Cf section II, 3e.

In practice the TR and the anti-TR arms are broken and choke-flange joints inserted between the T openings and the cavities. In the case of the 1B24 the joint is made at the window, and in the case of the 724B TR it is an effective $\lambda_g/4$ from the window. The anti-TR is usually designed to have a loss of 1.0 db when placed at the end of a line.

Testing methods for the TR box—(1) Tuning range.—Tuning range may be measured in two distinct ways: One, by measuring the transmission characteristics of the cavity; and two, by measuring the

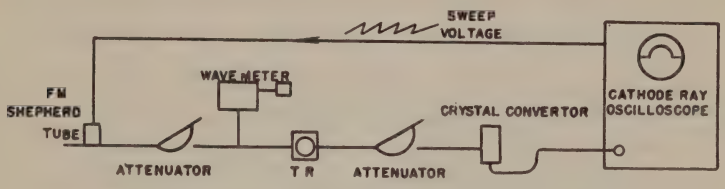


FIGURE 84.—F-m set up for TR tuning by transmission method.

input impedance with the output short-circuited (this amounts to measuring the absorption). The transmission may be measured in two ways. The method requiring the simplest equipment employs a variable oscillator feeding through a wave guide and absorption wave meter into a TR box and through the TR into a reasonably well-matched crystal detector connected to a milliammeter. Resonance is indicated by a peak in the crystal current. The wavelength is measured by tuning the wave meter until a dip occurs in the crystal current.

Another method of measuring the transmission range is with a frequency-modulated test bench set-up, as shown in figure 84.

When the TR is tuned to resonance the trace on the oscilloscope screen is as shown in figure 85.

FIGURE 85.—Transmission curve for TR as shown on oscilloscope.

The dotted line is the trace you get when the TR is replaced by a dummy section of guide. The wave meter can be tuned until the pip from it is at the peak of the transmission curve, and the wave length determined. This method is not accurate for a TR of Q less than 200 or so, unless only an approximate tuning range determination is wanted. When the TR is used as an absorption cavity, it is placed on the T arm of a $\lambda_g/4$ T section (E-type), and a shorting plate is placed across the flange on the output side of the TR, which should also be $\lambda_g/4$ from the output window of the TR. (See figure 86.)

With the TR at resonance the oscilloscope trace looks as in figure 87, and here again the wave meter pip can be brought into coincidence with the bottom of the dip to read the resonant wave length. This method is considerably more accurate than the other, and generally more convenient.

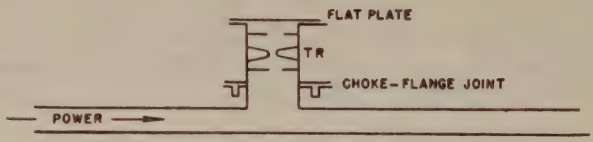


FIGURE 86.—Position of TR when tuned on an F-m system by absorption method.

It should be mentioned here that in any tuning or loss measurement on the TR the wide-range tuning screw should be tightly locked when measurements are being taken, and cannot be twiddled for the final adjustment, since it may introduce considerable loss when not tightly locked.

FIGURE 87.—Absorption curve as shown on oscilloscope.

(2) *Low-level loss, Q_L, Q_0.*—These are best made by standing wave measurements using a probe and slotted section. Methods for making these measurements are described in section IV, 10. The most convenient method is the reflection method which it might be well to review here. The set-up is as shown in figure 88.

FIGURE 88.—CW set-up for measuring the Q of a TR.

The procedure:

1. Match termination until the power standing wave ratio, r^2, is less than 1.05.

2. Tune the TR to resonance by minimizing the standing wave ratio in front of the TR.

3. Measure the maximum current, I_z, minimum current, I_n, in front of TR, and the current beyond, I', by moving the probe from one slotted section to the other. Measure λ_0.

4. Detune the oscillator until the power standing wave ratio is about 10 to 20 in front of the TR. Measure I_z, I_n, λ.

5. At resonance, $\beta^2 = I_x/I_n$, cavity loss $= 10 \log \frac{\sqrt{I_x I_n}}{I'}$, transmission $T = \frac{I'}{\sqrt{I_x I_n}}$.
6. Off resonance, $x^2 = I_x/I_n$.
7. Calculate $\Sigma = \left[\frac{\beta(x+1)^2}{x(\beta+1)^2} - 1\right]^{\frac{1}{2}}$,
8. Compute $Q_L = \frac{\lambda_0}{2(\lambda - \lambda_0)}\Sigma$, $Q_0 = \frac{2Q_L}{1-T}$*.

Note that the figure for cavity loss must be corrected as follows. The TR is replaced by a dummy section of guide of approximately the same length with a flange on each end. The current in front, I_1, and the current beyond the dummy, I_2, are measured. The apparent loss is 10 log I_1/I_2; this is subtracted from the previously computed TR loss to give the corrected loss. This correction takes account of losses in the couplings and differences between the slotted sections. One sometimes turns up with an apparent gain through the dummy because of these differences in slotted sections.

Note that the value of Q_0 should be constant for a given cavity, regardless of changes in the windows. This makes a good check on the measurements. These Q measurements are not as accurate as might be desired, and one can expect maximum deviations in the Q value of ± 10 percent, though experience will reduce this.

3. LOCAL OSCILLATORS AND MIXERS

a. Local Oscillators.

Description.—With the exception of the GL–446 Lighthouse tube, all 10 cm. local oscillators in common use at present are reflex velocity-modulation tubes. These tubes have a single resonant cavity, in which two grids or apertures are provided at a voltage maximum. An electron-emitting cathode is placed nearby at a negative potential, so that a beam of electrons is drawn from it and passes through the cavity, entering and leaving by the grids or apertures. As the electrons pass through the cavity, they are alternately speeded up and slowed down by the radio-frequency electric field in the cavity. A certain time after the electrons leave the cavity, the accelerated electrons catch up to the slowed electrons, causing a momentary concentration of electrons. The beam is now said to be "bunched."

In a reflex tube, the bunched beam is turned around by a highly negative electrode called the reflector or repeller, and pasees again through the cavity at such a time that the electron bunches are slowed down, and hence do work on the field in the cavity. This work

*This is only strictly true when the TR is matched to the line, and is approximately true for the case of equal coupling windows.

makes oscillation possible. See the accompanying diagram of a reflex Klystron (figure 89).

In a reflex tube the oscillating frequency is a function of the reflector voltage. For only one value of reflector voltage will the transit time of the electrons be correct for oscillation at the true resonant frequency of the cavity. At slightly different voltages the tube will not drop out of oscillation but will shift frequency to compensate. The total amount of frequency shift possible without losing oscillation varies from almost zero to about 100 mc. This depends on the type of tube, i. e., the resonator grid and reflector structure, the number of oscillation periods (or transit angle) between the time electrons just enter the grid structure and the time they are reflected to it, and the accelerating potential. In general the amount of electron tuning thus obtained increases with increasing accelerating voltage and decreasing reflector voltage. The variation of frequency with reflector voltage for a 707A tube operating under stated conditions is shown in figure 90.

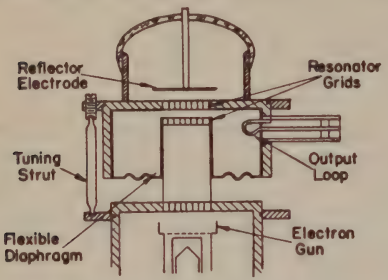


FIGURE 89.—Schematic view of a reflex Klystron, a typical velocity modulation tube.

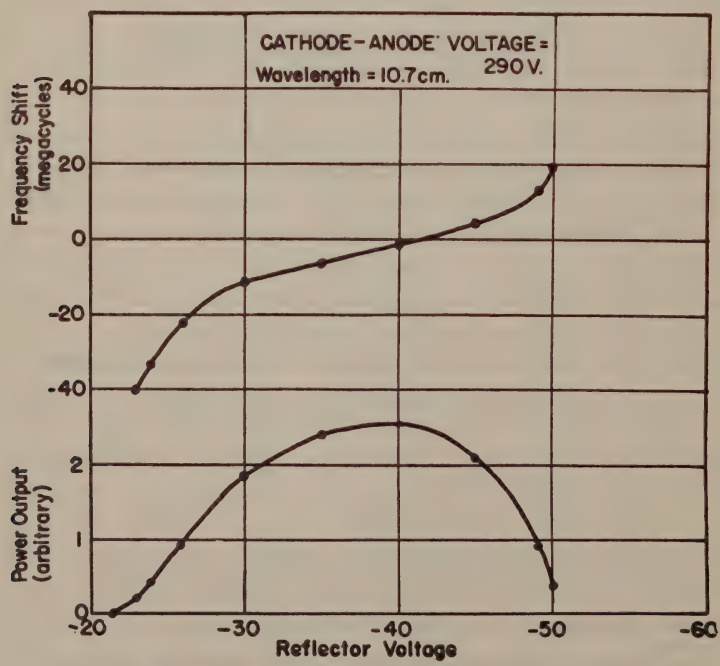


FIGURE 90.—Reflector characteristics of the 707A.

Characteristics of various local oscillators.—

707A (McNally tube).

Made by Western Electric, Raytheon, and Sylvania. Cavity is removable. A simple type of cavity is tunable by adjusting plugs over a wavelength range of 5 millimeters or more at any wavelength from 8 to 12 cm. Cathode to resonator voltage is 200 to 300 volts. Accelerating grid tied to resonator. Reflector voltage adjustable between 10 and 150 volts negative to cathode. Oscillation will occur at 2 or 3 points in this range. Variation of frequency with reflector voltage in any one mode, or operating condition, is large, about 30 megacycles between reflector voltages at which power output is reduced to one-half of maximum. The appearance of the 707A is shown in figure 91.

707B. Temperature compensated model of 707A.

2K28. Similar to 707B but smaller glass structure. Will fit same cavity as 707A or 707B.

726A. Made by Western Electric in moderate quantity. A compact, all-metal, version of the 707A. Reserved for those applications where small size and weight are of utmost importance. Tunable by adjustment of a screw from 8.9 to 9.4 cm. Exterior is identical with 723A except for slight difference in antenna structure.

GL–446 (Lighthouse tube). A small triode, made by General Electric and R. C. A. When used in a suitable cavity combination, oscillates at wavelengths of 11 cm. or longer. Selected tubes will oscillate in the 9 to 10 cm. range. Cathode to anode voltage 150 to 250 volts. Grid grounded. Cathode biased positive by grid leak resistor. (See figure 91).

723A/B (Shepherd-Pierce tube). This is the only 3 cm. local oscillator now in production. A compact, all metal, velocity modulation tube. Applied voltages same as 707A. Tunable by adjustment of a screw from 3.14 to 3.445 cm. Coaxial output line projects as a probe into waveguide. Figure 92 is a photograph of the 723A/B and its waveguide mount.

2K25. Made by Western Electric. Same as 723A/B except that tuner has been designed to cover a greater wavelength range. The range is approximately 3.1 to 3.6 centimeters.

417, 417A. Made by Sperry and Westinghouse. Cavity is integral part of tube. Tunable from 9 to 11 cm. by adjusting screws. Cathode to resonator voltage 400 to 600 volts. Control grid voltage 15 to 25 volts positive to cathode. Reflector adjustable between 0 and 100 volts negative to cathode. Oscillation will usually occur at two points in this range. Variation of frequency with reflector voltage is small (5 megacycles). Coaxial output line. The tube is illustrated in figure 92.

Power supplies.—Well-regulated power supplies are required for the velocity modulation tubes, to stabilize the frequency sufficiently for use as local oscillators. The cathode to anode supply drain will be 20–30 ma.; the reflector supply drain is several microamperes for the

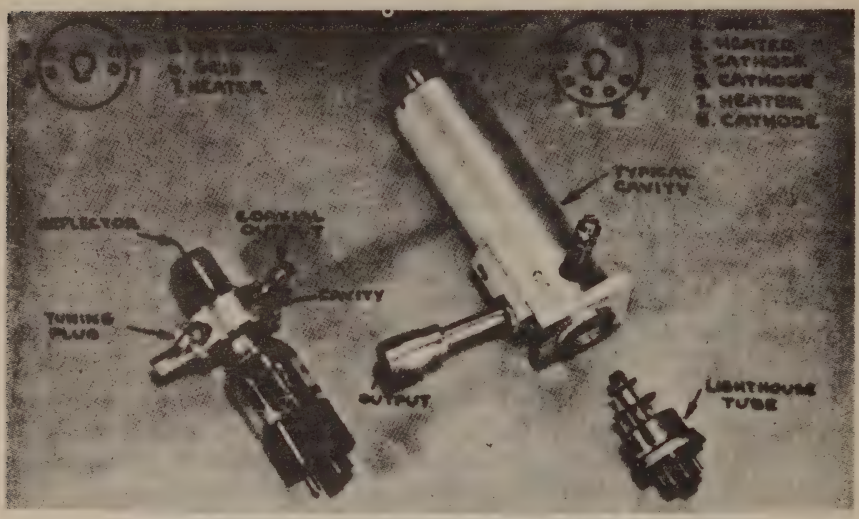


FIGURE 91.—Local oscillators and base connections. Left, the 707A or McNally tube. Right, the GL–446 or Lighthouse tube.

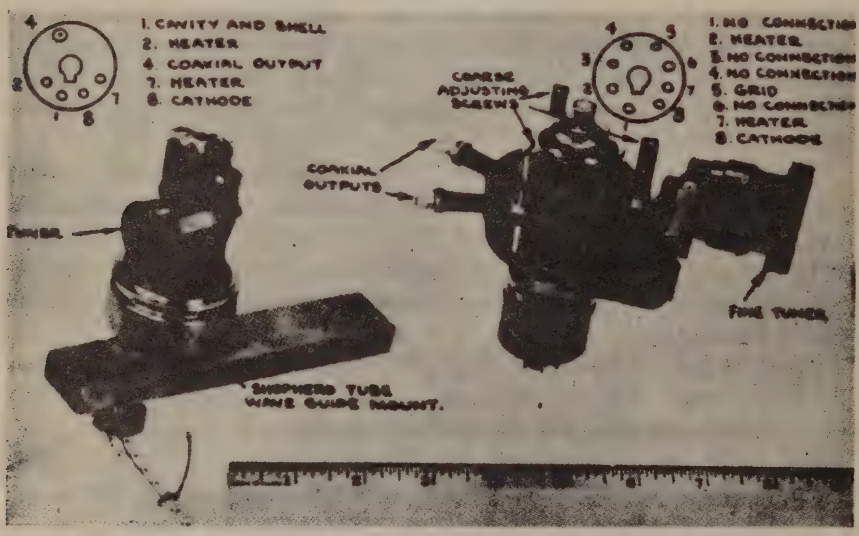


FIGURE 92.—Local oscillators and base connections. Left, the 723A/B or Shepherd tube. Right, the 417–A or reflex Klystron.

707A, 726A, 723A/B, and [illegible] for the 417. The voltages are usually obtained from supplies already available in the receiver or other components. The voltage is stabilized by an electronic regulator or by

VR (gas) tubes. The reflector voltage is about four times as critical as the cathode to anode voltage. Approximate frequency versus reflector voltage sensitivity values are listed below:

Tube	$\frac{\text{mc.}}{\text{volt}}$ on reflector
417	0.16
707A	1.1
723A/B	2.3

Notes on operation.—The normal power output is about 50 milliwatts for the 10 cm. tubes, and 25 for the 723A/B. About 2 milliwatts needed in the crystal mixer; hence loose coupling is used between the local oscillator and crystal mixer. This loose coupling also insures that little of the received signal is absorbed by the local oscillator cavity and line.

A fine control local oscillator tuning adjustment is usually provided the operator. On the 417 and GL–446 this knob is mechanically coupled to the cavity; on the 707A, 726A, and 723A/B it is a potentiometer controlling the reflector voltage.

In systems employing automatic frequency control, an error voltage is fed onto the reflector in such a way as to keep the local oscillator properly tuned with respect to the transmitter. The 417 and GL–446 tubes cannot be used in this application because their electrical tuning range is inadequate.

b. Mixers.

Heterodyne detection.—In microwave radar, the transmitted frequency is so high that only recently has a good receiver been built which amplifies the received signal at radio frequency. Hence, it is customary to convert the signal to a lower frequency before it is amplified. There are two methods of doing this. The most direct approach is to rectify the signal and change it to a video pulse which may be amplified. This is done for some purposes. However, in the majority of cases it is more convenient to change the frequency of the pulse to some intermediate frequency, and first amplify the intermediate frequency, and change it to a video pulse after a large amplification. There are three main reasons for this: (1) It is possible to detect smaller signals; (2) it is possible to obtain more selectivity; and (3) it is somewhat more convenient to have the gain of the receiver distributed between two frequencies (intermediate and video) so as to avoid trouble from feedback.

Let us consider the question of a figure of merit for the sensitivity of a receiver. It is well known that if an amplifier has enough gain, a random voltage appears at the output terminals. This random voltage is called "noise" and arises in several ways, If we had a

perfect amplifier, the noise would come from the random motion of electrons in the grid circuit of the first amplifier tube. Radio tubes all contribute noise because of fluctuations in emission, interception noise at grids, and shot noise at the plate. For any receiver, there will be a definite amount of input signal power which is needed to make the output signal power equal to the output noise power. A perfect receiver requires kTB watts (k=Boltzmann's constant; T= temperature in degrees Kelvin; B=band width in c. p. s.). The *noise figure* of a receiver is the figure of merit for sensitivity and is defined as the ratio of the input power to kTB when the output signal power equals the output noise power. Normally the noise figure is expressed in db.

Ordinary heterodyne radio receivers use tubes for first detectors. Many tubes have been tried for the same purpose in microwave reception but so far all have proved quite noisy. The Westinghouse Electric Co. 708A and British diode CV58 are the only tubes in use. At the 10 cm. band tubes in service seem to be about 25 db. worse than kTB, while the best has been recorded at 16 db. The noise situation is even worse for reception in the 3 cm. band. As a result attention has been turned to the crystal detector, a device once widely used at ordinary radio frequencies but for many years completely superseded in this field by tubes. This relies on rectification at the barrier between a semiconductor (crystal) and sharpened metal wire ("whisker").

The actual housing of the crystal is called the "mixer." Several such have been designed for crystals in coax, in the 10 cm. band. Such devices must have inputs for signal and local oscillators and an i-f output. There must also be provision for bypassing the r-f and a filter to prevent its escape through the i–f line.

10 cm. mixers.—The earliest of these devices at 10 cm. was the "pot" mixer, as illustrated in figures 93 and 96 (left). Signal and local oscillator power are fed into the center post by adjustable probes inserted through the diametrically opposed apertures. The crystal is fitted in through the third hub as shown and held in place with a cap. The large cover on the "pot" can be screwed in or out to afford an adjustable capacity for tuning purpose. The position of the signal probe, always tightly coupled, can be varied to give a second tuning adjustment. The local oscillator probe is loosely coupled and can be adjusted to give required crystal current, usually 0.6 ma. The intermediate frequency output and the direct crystal current are drawn off through the lead at the bottom to the preamplifier.

Another type is the "pencil" mixer, which couples with a loop directly into the TR box. The coupling can be varied by rotation of the loop. An adjustable probe feeds in local oscillator power from

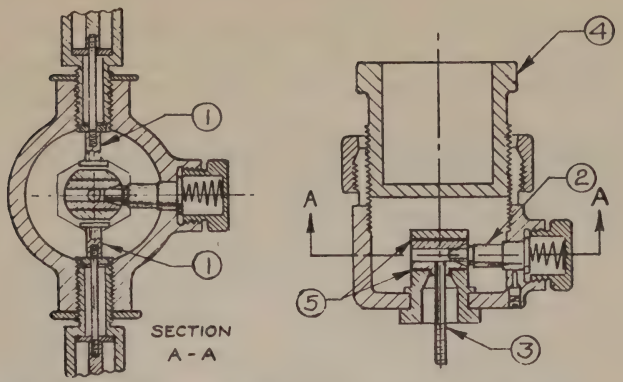


FIGURE 93.—10 cm. crystal mixers—(A) the Pot mixer. ① inputs for local oscillator and signal, ② crystal cartridge, ③ output for intermediate frequency, ④ adjustable pot top, and ⑤ insulator for r-f capacity bypass.

the side. See figures 94 and 97 (right). The piece which holds the base of the crystal in a finger grip is separated from the outer conductor of the line by a polystyrene sleeve which acts as a choke for radio frequencies.

Because of the general advisability of having as few adjustments and as much "preplumbing" as possible, there has been considerable interest directed toward a fixed-tuned mixer. The use of a fixed-tuned mixer in systems is highly advisable for several reasons. A tunable mixer is quite difficult to adjust for best results. Any device which has tuning adjustment will undoubtedly be improperly adjusted at least part of the time. The most important point however is that the tuning of such a mixer must necessarily present a large range of impedances to the output of the TR box. Since the leakage power of a TR box is highly dependent upon the output impedance presented to it, this process is very apt to result in burned out crystals.

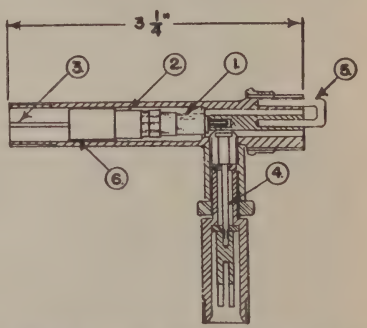


FIGURE 94.—10 cm. crystal mixers—(B) the Pencil mixer. ① crystal, ② adapter (finger grip on crystal), ③ intermediate frequency output, ④ local oscillator feed, ⑤ coupling loop into TR box, and ⑥ capacity short for r-f.

Since maximum receiver sensitivity results when the received signal is matched into the crystal, to make a fixed-tuned mixer the spread of crystal impedances must be fairly small. This is indeed found to be the case especially since production tests have been introduced which require the crystal to have at least a given sensitivity when used in a fixed-tuned mixer. Thus those which reflect abnormal amount of power are eliminated. These points make the fixed-tuned mixer practicable.

Thus the use of a fixed-tuned mixer eliminates the objections raised to a tunable mixer and limits the possible range of impedances presented to the TR box to reasonable values and so increases the life expectancy of crystals. The Pound fixed-tuned 10 cm. mixer is shown in figures 95 and 97 (left). This again couples directly with a loop into the TR box. The only adjustment is that of the local oscillator which is fed in by a probe in a sleeve. The double choke to bypass r-f and prevent its escape along the i-f lead is clearly visible. It has type N connectors. The Pound mixer is fixed tuned over the 9 to 11 cm wavelength range.

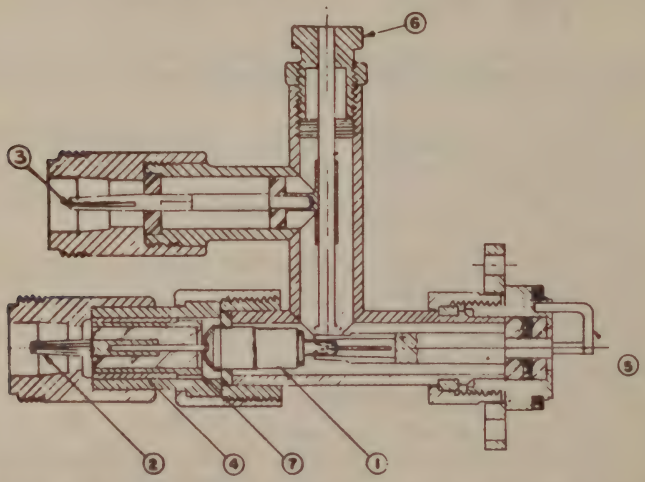


FIGURE 95.—10 cm. crystal mixers—(C) the Pound mixer. ① crystal, ③ intermediate frequency output, ④ local oscillator feed, ⑥ capacity short for r-f (insulator), ⑤ coupling loop into TR box, ⑧ knob for adjusting local oscillator drive, ⑦ cup for r-f filter.

Another fixed-tuned mixer was designed by W. M. Sharpless of the Bell Laboratories primarily for uniformity in testing. It has Holmdel (D–162143) fittings. All sliding adjustments can be held tight by set screws as shown. In this mixer a crystal can be plugged in and out directly. It is shown in figure 96 (right). The Sharpless mixer is not fixed tuned in the same sense as the above. It is tuned up and fixed for the average crystal at a particular wavelength and must be retuned when even small changes are made in the wavelength.

3 cm. mixers.—3 cm. wave guide mixers are shown in figure 98. The mixer proper consists of a piece of wave guide across which is plugged a crystal parallel to the electric field. In the tunable mixer the crystal is matched into the guide by means of an adjustable plunger behind it and one of two tuning screws in front of it. See figure 98 (right).

All of the arguments for fixed-tuned mixers at 10 cm. also apply in the 3 cm. region. Making a fixed-tuned mixer here, however, is more

difficult because the spread of crystal impedances is greater than in the 10 cm. band. Since in this band the crystal is shunted across the guide, we usually speak of admittances and it is found that the spread of susceptances is large and of conductances is tolerable. Fixed

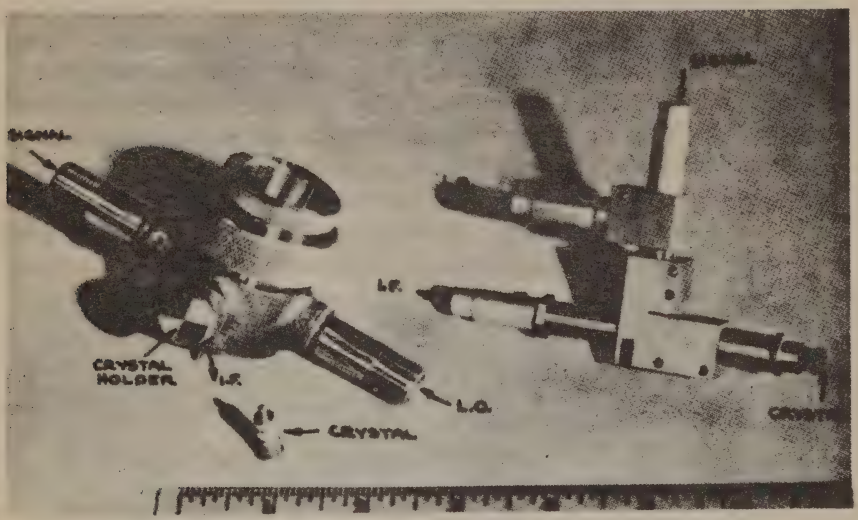


FIGURE 96.—10 cm. mixers. Left, the pot mixer. Right, the Sharpless mixer.

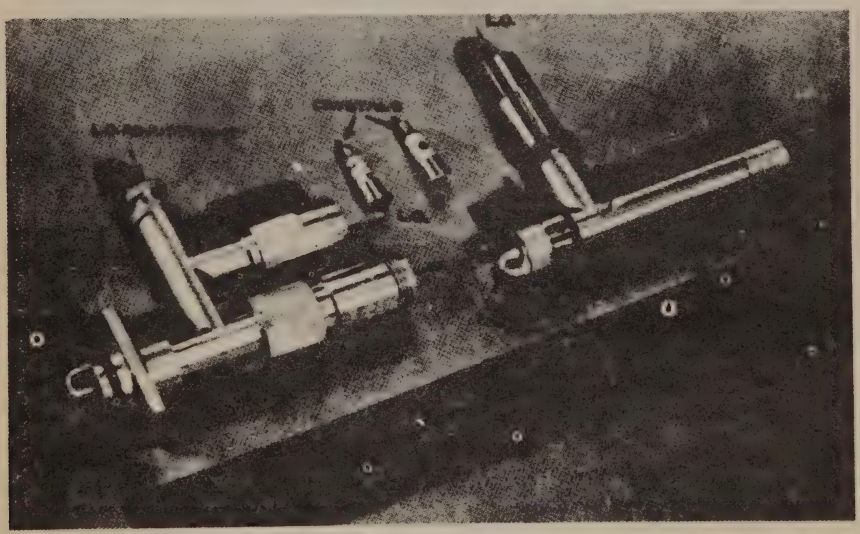


FIGURE 97.—10 cm. mixers. Left, the Pound mixer. Right, the pencil mixer.

matching devices to match the average crystal to the guide are incorporated. The crystal is so placed relative to the TR that the TR is capable of tuning out the susceptance of the crystal when tuned for maximum signal. This is true because the leakage power from a TR

is essentially independent of the susceptance of the load and dependent only on the conductance. By this means a mixer which has all of the virtues of a fixed-tuned mixer is produced and yet it accepts a wide range of crystal admittances.

The normal spread of conductances of crystals results in about 1 db. signal loss in the most unfavorable cases in both 3 cm. and 10 cm. band fixed-tuned mixers, but it must be remembered that this is not the average case. In the average case the fixed-tuned mixer is equivalent to a tunable mixer when the latter is properly tuned, which is probably not always true. Furthermore it is quite inadvisable to tune such a mixer for maximum signals on the system since crystal burn-out is imminent in such a procedure. This necessitates a test bench tune-up of the tunable variety of mixer and this is a very in-

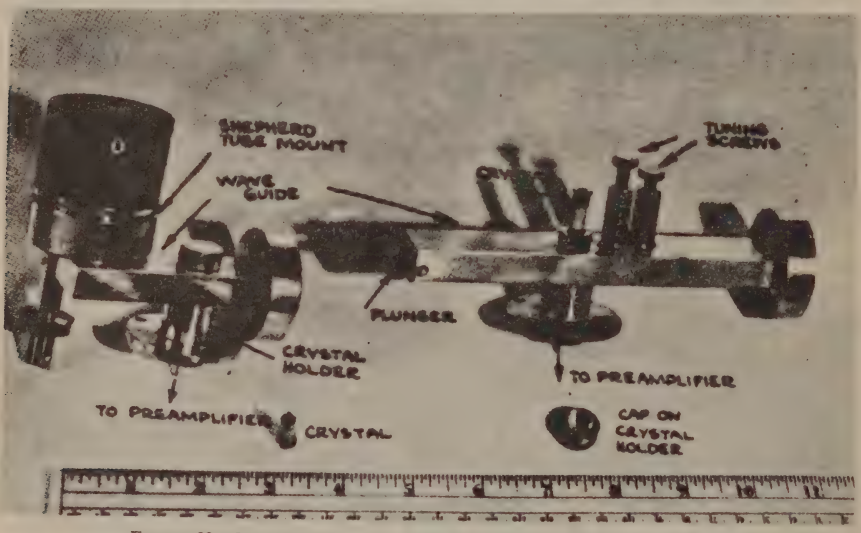


FIGURE 98.—3 cm. mixers. Left, fixed tuned mixer. Right, tunable mixer.

convenient thing for field operation. A 3 cm. fixed-tuned mixer is shown in figure 98 (left).

Recently X band double mixers have come into use. These include a second mixer for automatic frequency control of signals. Here the local oscillator is tightly coupled to a wave guide and small amounts of local oscillator power are taken off through windows to each of two fixed-tuned mixers.

c. Crystals.

Cartridge-type crystal converters were developed for microwave reception because a crystal rectifier is highly efficient as a first detector in a superheterodyne receiver for these frequencies. A sharpened tungsten whisker, placed on a smooth silicon surface, supplies the rectifying contact. For the most sensitive detection the contact area must be small to minimize capacity effects. In practice it is

of the order of 10^{-3} cm. in diameter. All present production crystal cartridges employ silicon for the semi-conductor but there is in progress some research investigating the possibilities of germanium for this role. The d–c characteristic of a typical unit is as shown in figure 99.

The most usual direction of electron flow is from metal to silicon and is called "p" type. The conduction through the silicon depends on impurities of the order of 0.1 of a percent and by a varying choice of the "doping" agent the rectification could be greatly changed or even reversed. Naturally the treatment of the silicon and the preparation of the surface is very critical in converter performance. In the actual process of assembly, i. e., pressing down the sharpened whisker on the silicon, the d–c characteristic of the crystal which is viewed on an oscilloscope, can often be improved by judicious "tapping" of the cartridge when done by a skilled operator. (Such treatment is definitely not recommended for the field.)

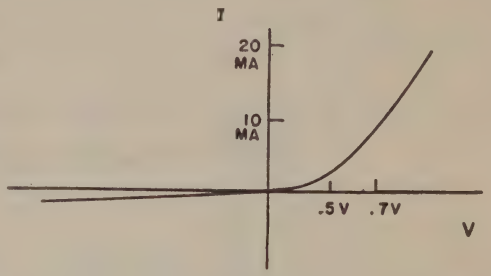


FIGURE 99.—Direct-current characteristics of a typical crystal.

In the original British Thomson Huston (B. T. H.) cartridge the silicon was placed in the pin end and the whisker was put in the base. The first American models copied the British closely. However, production in this country was standardized on a reverse polarity with the crystal in the base end, in which case the pin end of the crystal has positive polarity. The two largest producers at present are Western Electric Co. and Sylvania Electric Products, Inc. Drawings of their models are shown in figure 100. The external cartridge dimensions and tolerances have been standardized by the Navy. It can be seen from the outline drawings that the essential difference in the two types of cartridge lies in the method of adjusting the rectifying contact. This is adjusted in the Sylvania unit by advancing a screw in the base end of the crystal, on which is mounted the silicon. The Western Electric unit is adjusted by means of a screw in the pin end, which advances the sliding post on which the whisker is mounted. In the latter case adjustment is made without any relative rotation of whisker and silicon. After adjustment, the cartridge is impregnated with a filler the purpose of which is to increase the mechanical stability of the contact and to render the unit impervious to moisture.

A good d–c characteristic is not a sufficient criterion for crystal performance in the microwave region. An equivalent circuit of the rectifying contact is shown in figure 101. The contact itself is equiv-

alent to a resistance R_c in parallel with the capacity C of the boundary layer. In series with this element is the bulk resistance of the silicon R_f. For backward voltages R_c is very large, while for reasonably large forward voltages R_c is very small. If the frequency is high enough, the capacity C will shunt out the resistance of the contact in the backward direction. Using this picture Bethe has shown

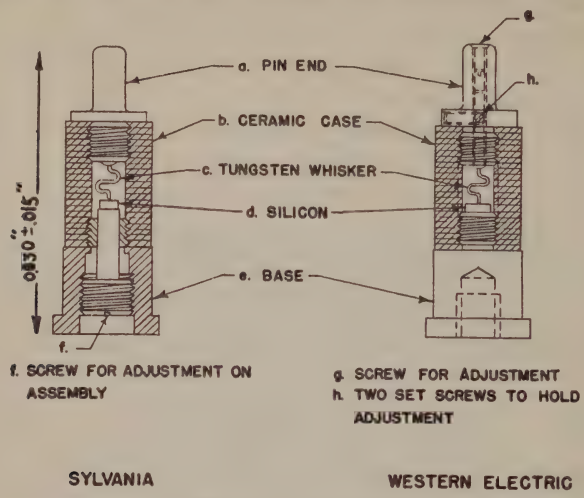


FIGURE 100.—Crystal cartridges.

that for high conversion efficiency at high frequencies the product CR_f must be small. The excellence of performance of a crystal used as a frequency converter will be determined primarily by the conversion loss of the crystal and by the noise which it introduces into the receiver.

It has been found experimentally that the power from the local oscillator causes the crystal to generate more noise than an equivalent

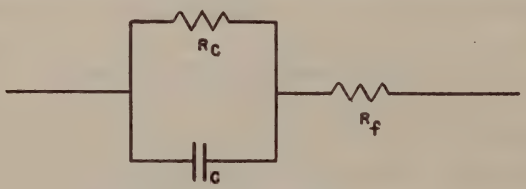


FIGURE 101.—Equivalent circuit of a crystal.

resistor at room temperature. The available noise power from a resistor commonly called Johnson noise, is given by kTB, where k in Boltzmann's constant, T is the absolute temperature, and B is the effective band width of frequency. The output noise ratio of the crystal N_r is then defined as the ratio of the available output noise power of the crystal to kTB. N_r is frequently called the "noise

temperature." This terminology arises from the fact that N_r is, by definition, the ratio T'/T where T' is the temperature at which the equivalent resistor would generate just as much noise as the crystal, and T is absolute room temperature.

A weak signal from the antenna becomes still weaker when it has become converted by the crystal to the intermediate frequency. The conversion loss L is defined as the ratio of the available r–f power output to the available i–f power input, and is always greater than unity.

The conversion loss and output noise ratio of the crystal are related to the noise figure of the receiver by the expression:

$$\overline{NF}_{rec}=L(N_r+\overline{NF}_{i-f}-1)$$

where $\overline{NF}_{rec}$ is the noise figure of the receiver and $\overline{NF}_{i-f}$ is the noise figure of the i-f amplifier. It has been found that L and N_r vary so widely from crystal to crystal that it is necessary to measure each unit and accept it for use only if it falls within specified limits.

Impairment of crystals in systems is usually caused by preignition transmission through the TR. This so-called spike is of very short duration, probably less than 0.01 microsecond. In such a short time the crystal and whisker are unable to conduct the dissipated heat away from the contact and the resulting temperature is probably proportional to the energy in the spike rather than its peak power. For this reason most crystals are now given a burn-out proof test by the manufacturer which involves exposing them to short d–c pulses (2.5×10^{-9} sec.) generated by the discharge of a coaxial line. This exposure is made before any measurements of performance. The proof test is specified in terms of the voltage applied to a specified coaxial line, from which the corresponding pulse energy can be calculated.

The table below lists the electrical properties of the crystal types specified for converter use, as of 1 May 1944.

Type	Max. L (db)	Max. N_r (times)	$\overline{NF}_{rec}$ (db) $\overline{NF}_{i-f}=5$ db	Burn-out pulse energy (ergs)	Band
1N21A	7.5	3.0	14.6	0.3	S
1N21B	6.0	2.0	12.2	2.0	S
1N23	10.0	3.0	17.1	0.3	X
1N23A	8.0	2.0	14.2	1.0	X
1N23B	6.5	2.0	12.7	0.3	X

Both the conversion loss and the output noise ratio are functions of the rectified crystal current which in turn is determined by the local oscillator level. Typical curves showing this dependence for a crystal in the 3-cm. band are shown in figure 102 (A). A curve showing the calculated value of the noise figure of two receivers using this particular

crystal is shown in figure 102 (B). The output noise ratio is not linear for all crystals, and the shape of the conversion loss curve varies somewhat from crystal to crystal, but the general shapes and trends of curves are the same. The minimum of the curve for the receiver noise

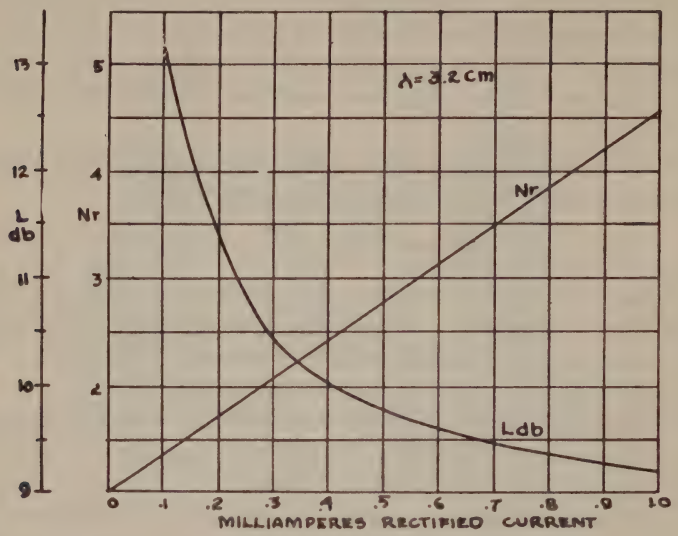


FIGURE 102 (A).—Graph showing conversion loss and noise ratio versus rectified current for a crystal in the 3-cm. band.

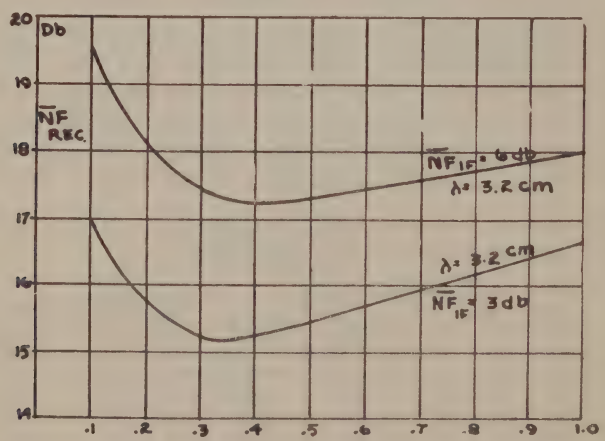


FIGURE 102 (B).—Graph showing the calculated value of the noise figure for two receivers using the crystal of figure 102 (A).

figure varies from crystal to crystal in the region from 0.3 to 0.8 ma. rectified current. In the 10-cm. band similar characteristics are found.

The effect of the application of d-c bias on crystal performance has

been noted by many investigators. It is found that negative bias is undesirable in that noise output increases and the signal is reduced. An improvement is obtained for small positive bias. This improvement varies widely from crystal to crystal and may be as much as 1 db. in the noise factor of a receiver having $\overline{NF}_{i-f}=5$ db. The improvement occurs mainly in the reduction of output noise and is most evident in noisy units; there is usually an improvement in the signal output as well. The most favorable bias appears to be in the neighborhood of +0.10 volt.

The i-f impedance at 30 megacycles can be measured by means of an impedance bridge while the crystal is being excited by a local oscillator. The impedance is found to vary with rectified crystal current, with the r-f tuning, and with the mixed used. With r-f

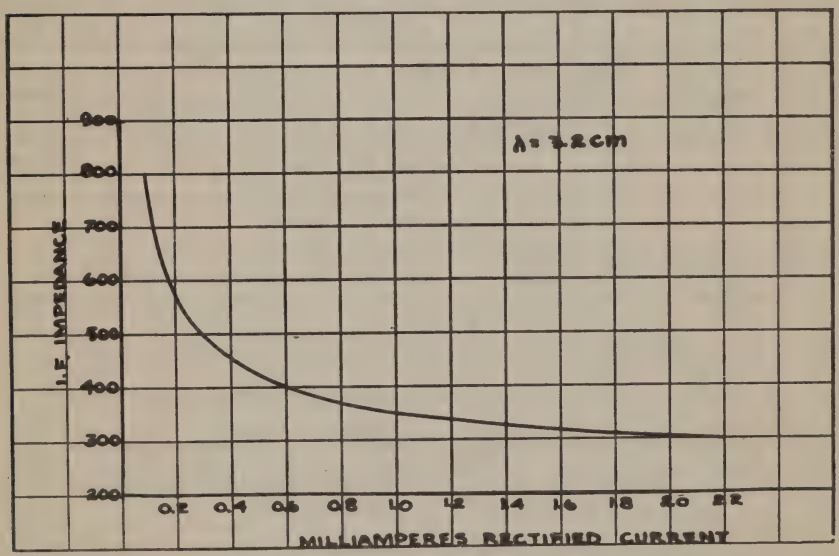


FIGURE 103.—Change of intermediate frequency impedance with rectified crystal current for a typical crystal.

match and a rectified current of 0.6 ma. the i-f impedance varies from crystal to crystal in the range from 150 to 600 ohms, most crystals ranging from 200 to 400 ohms. The average value is in the neighborhood of 300 ohms. The i-f impedance may be determined by a d-c method if one assumed that a crystal excited by a local oscillator is equivalent to a d-c generator having an internal resistance equal to its i-f resistance. The magnitude of the external resistance required to reduce the rectified crystal current to half its value under short circuit conditions is then equal to the crystal impedance.

At low r-f power levels crystals act as square law detectors, that is the rectified crystal current is proportional to the r-f power. The

upper limit of the region in which the crystal is accurately square law varies from crystal to crystal, but is usually found in the region from 2 to 10 microwatts. The graph of figure 104, obtained with a rather sensitive crystal, shows the rectified current and d-c resistance as a function of r-f power in the 10-cm. band. The d-c resistance in this region varies from crystal to crystal between limits of approximately 1,000 ohms to 15,000 ohms.

There is no evidence that crystals have a finite shelf life, although day-to-day measurements indicate that small fluctuations occur in some crystals. Experience has shown that crystals may be seriously damaged in several ways by improper handling.

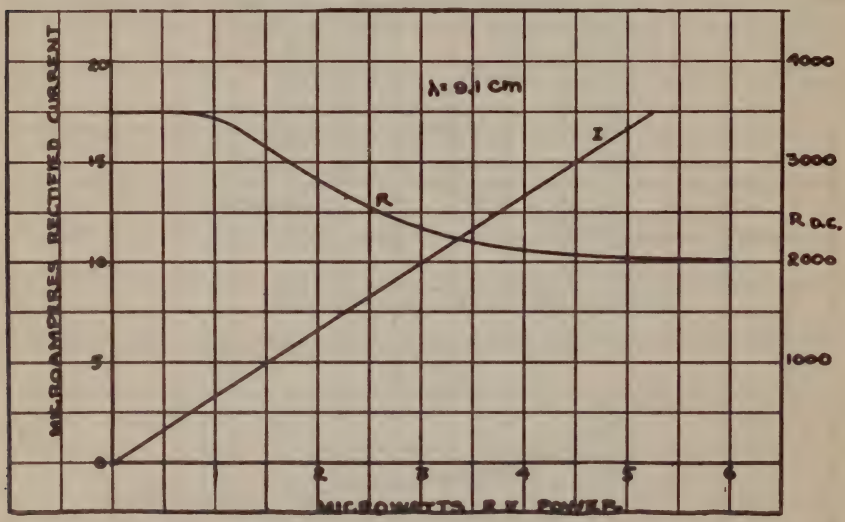


FIGURE 104.—Graph showing rectified crystal current and direct-current resistance as a function of radio frequency power in the 10-cm. band.

In the first place, static charge may be easily accumulated by the operator, particularly in a dry atmosphere. In this case there exists the possibility of damage to the crystal due to the passage of discharge currents when the cartridge is inserted in the apparatus. Minute static charges not readily detectable are sufficient to cause damage. Such impairment can be avoided by holding the cartridge by the base and making electrical contact with the equipment through the hand while inserting the cartridge. Crystals may also be damaged by a discharge to ground through the body of the operator when the crystal is inserted in apparatus which is operating at other than ground potential.

In the second place, crystals may also be damaged by voltages in connecting wires, induced by neighboring electrical apparatus. In this way voltage shocks may be delivered to the crystal by the opening

and closing of electrical circuits or by pick-up of video power. Such damage may be avoided by careful shielding of the connecting leads.

A third source of damage lies in the exposure of crystal cartridges to intense r-f fields. This can be avoided by storing the cartridges in metal containers or wrapping them in metal foil when they are not in use.

Finally, crystal cartridges will stand only a limited amount of mechanical shock and should be handled with the same care as vacuum tubes.

Knowledge of changes in the efficiency of a crystal as a converter can only be obtained with certainty by measurements of conversion loss and noise temperature. However, measurements of the d-c back resistances is a useful field test for crystal stability. Nearly always a decrease in the efficiency of the converter is accompanied by a sizable drop in back resistance, although there seems to be no correlation between the magnitudes of the changes; in fact it is occasionally observed that the back resistance may change appreciably without an appreciable change in noise or signal output. This test should, therefore, only be used if equipment is not available for measuring conversion loss and noise temperature changes.

4. ANTENNA MATCHING

The impedance of an antenna feed is a function of: (1) the wave length, (2) the geometry of the feed and the feed line inside of the parabolic dish, (3) the size and focal length of the dish, (4) the position of the feed in the dish, and (5) reflection from nearby objects. The position of the antenna in a dish is fixed by considerations of the antenna pattern (a suitable compromise between maximum gain and minimum side lobes). Associated with an antenna feed is the center of feed; namely, the apparent point source. This point is placed in a dish so that it is an integral number of half wavelengths from the vertex and as near the focal point as possible, in order to obtain the best gain.

The actual matching of an antenna to the r-f line should be done with no obstructions in the view of the dish. Pointing out of an open window with no nearby buildings is desirable. It is also possible to match an antenna in a room if suitable precautions are taken. The standard technique of designing matching transformers is used; namely, determination of the size and position of a transformer from measurements of the standing wave ratio in voltage and position of the maximum and minimum voltage points. For coaxial lines, a quarter-wavelength transformer on the inner conductor is used. However, it is sometimes necessary to use other than quarter-wavelength transformers if conditions of space are limiting factors. In the case of

rectangular or round wave guide, the matching is done by inserting an inductive window at the proper place. Transformers are generally placed as close to the antenna feed as possible in order not to increase the frequency sensitivity of the antenna. A properly designed antenna feed should have a standing wave ratio less than 2.5 in voltage before the final matching step is made. Most antennas are sufficiently broad-banded, so that after matching, the standing wave ratio will be less than 1.2 for a frequency spread of ± 1 percent about the spot frequency.

There are several interesting properties of matched antennas (matched at the half-wave point nearest the focus of some dish). (1) If this matched antenna is put into another dish of different size or focal length, and if the center of feed is placed at the half-wave point nearest the focus of this new dish, then the mismatch will be a minimum. The actual amount of mismatch will depend on the focal length. (2) If, after matching, the center of feed to vertex distance is varied the increase in mismatch will be small for a small movement. Maximum mismatch will occur where a displacement of $\pm\lambda/4$ is made. The smaller the dish the more rapid the increase of mismatch. (3) Small angular displacements of the feed line from the axis of the dish have very little effect on the match. This displacement can be as great as $+6°$ in a 30-inch x 10.6-inch f (i. e. focal length) dish with only a change in mismatch from 1.0 to 1.2. (4) It is important to note that the presence of a turret about a dish will change the apparent impedance of the antenna. A well-designed turret should not increase the mismatch by more than a factor of 1.2 and the shift in position of a maximum voltage point should not be changed by more than $\pm\lambda/30$.

5. R–F HEADS

In many radar installations the radio-frequency line connecting the equipment to the antenna would be long and difficult to install if the radio-frequency components were located with the modulator and power equipment. For example, the installation of a wave guide connecting an antenna located in an airplane wing with radar equipment in the plane cabin is almost impossible due to the large number of wing struts the guide must avoid. In addition, a typical installation of wave guide 30 feet long has 2 to 3 db loss at best.

One way to eliminate the long radio-frequency line is to locate the transmitter components and certain auxiliary equipment close to the antenna, and then supply the energy to the transmitter by means of a high-voltage cable. Figure 105 shows an r-f head developed for use with a 3-cm. band ASV set. This set is designed to be mounted behind the spinner in a nacelle mounted on an airplane wing.

Figure 106 is a block diagram of the r-f head shown in figure 105. This set, which is typical of many small assemblies, contains the copmonents listed in the figure caption. Power for the preamplifier

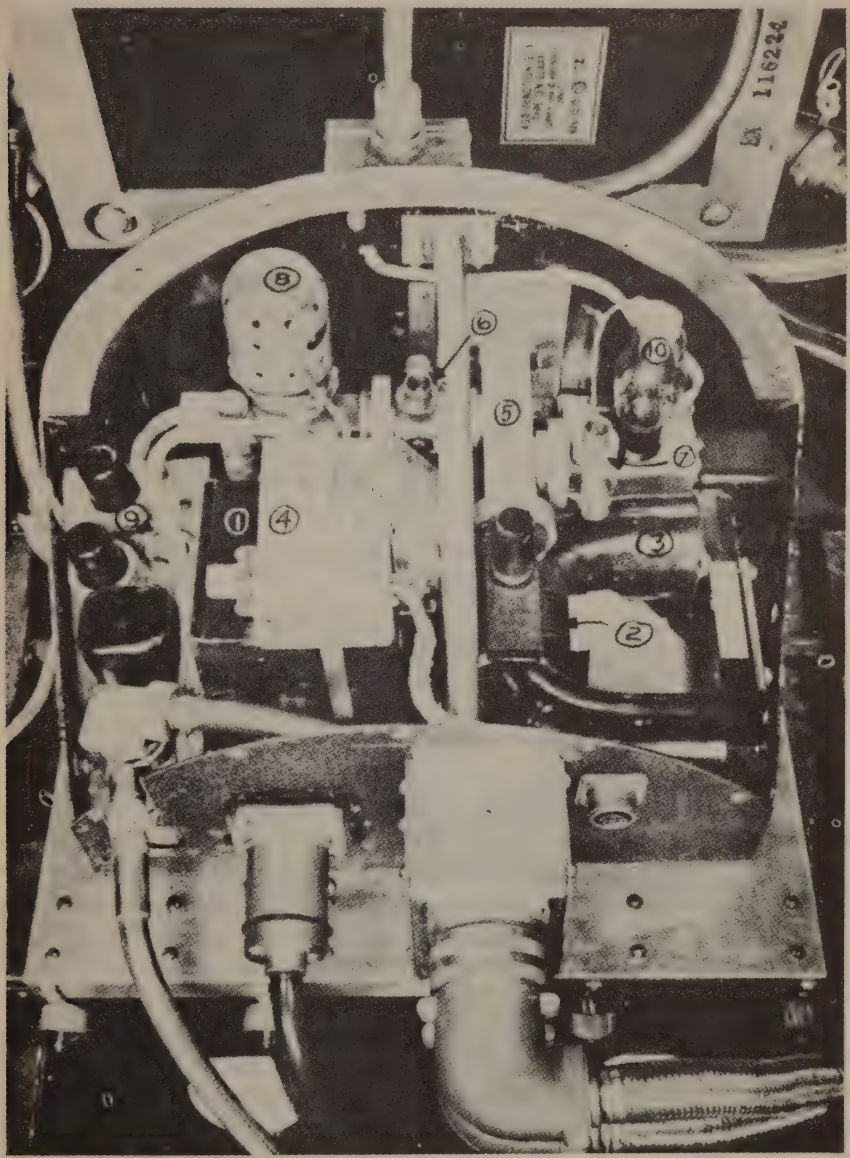


FIGURE 105.—R-f head. 3 cm. (See figure 106 for identification of number designations.)

and local oscillator is furnished to the head by a cable connecting with the receiver in the cabin. The intermediate-frequency output of the preamplifier is fed to the receiver by a separate low-loss cable.

Because of the limited space available in the wing of the planes for which the set described above was designed the unit was not pressurized; however, in some designs where more space was available for installation the entire unit was mounted in a pressurized can, a practice which is deemed necessary at the present time for installations that are planned for extremely high-altitude flying.

R-f heads are made for both 10- and 3-cm. band operation and find use in shipboard and ground station equipment as well as in aircraft units.

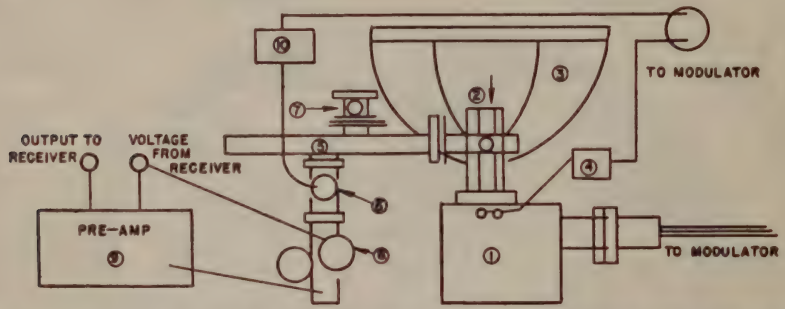


FIGURE 106.—Block diagram of a 3-cm. r-f head. ① the pulse transformer, which receives power from the modulator at 3,000 volts by means of the high voltage pulse cable. The transformer output feeds the magnetron 11 kilovolts. ③ the magnetron, which plugs into a well in the pulse transformer, ③ magnet, ④ magnetron filament transformer, ⑤ duplexing assembly, ⑥ TR cavity, ⑦ anti-TR cavity, ⑧ mixer and two local oscillator mounts, ⑨ preamplifier, ⑩ keep-alive supply.

6. PRESSURIZING

The purpose of pressurizing r-f lines is twofold: (1) To prevent break-down of the components at high altitudes and (2) to protect against transmission losses caused by materials in the atmosphere, such as dirt and water.

Since a large number of radar systems are to be used in high-altitude fighters and bombers, some protection must be provided for components which tend to break down or have corona discharge at low air densities. In the 3-cm. band most of the components will not cause break-down at the altitudes and magnetron powers now available. However, with the advent of higher-altitude planes and more efficient magnetrons it has become necessary to pressurize many systems. Theoretical considerations and actual measurements indicate that the 3-cm. band wave guide will break down at about 1 megawatt at atmospheric pressure, but other components in an r-f line, such as rotary joints and matching devices, will break down at much lower values. The weakest link in the 3-cm. r-f chain at present is the magnetron coupler. If typical values are desired, we can say that at present this coupler will break down at 15 kw. at 20,000 feet. These values imply a break-down power of about 60 kw. on the ground, which has been verified experimentally. Figure 107 shows how the

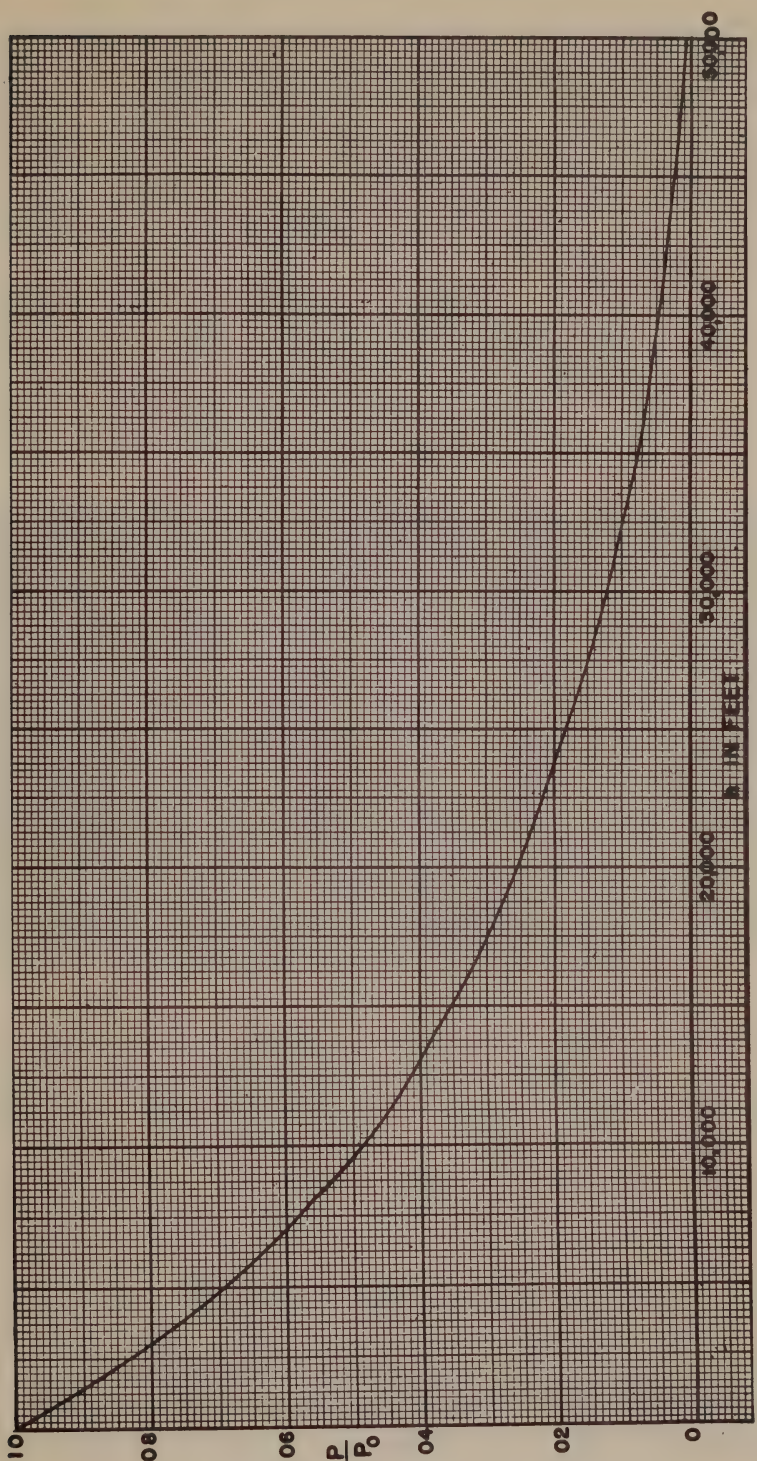


FIGURE 107.—Break-down power as a function of height.

break-down power (P) theoretically varies as a function of the altitude. P_0 is the break-down power at atmospheric pressure.

The second reason for pressurizing is at present the more important. On ship, ground, and plane installations, dirt and water are likely to accumulate in the line. Although this accumulation will not affect break-down to any considerable extent, it will certainly increase the transmission loss. Pressurizing is the means of keeping this unwanted material outside the lines.

Usually the complete r-f line is pressurized, from the magnetron to the antenna. The various joints in both coaxial lines and wave guides are pressurized by means of a rubber gasket placed between the two halves of each connector (figure 108). The antenna is pressurized by means of a polystyrene or lucite (plexiglass, methyl methacrylate) cup which surrounds the dipole (figure 109). On some sys-

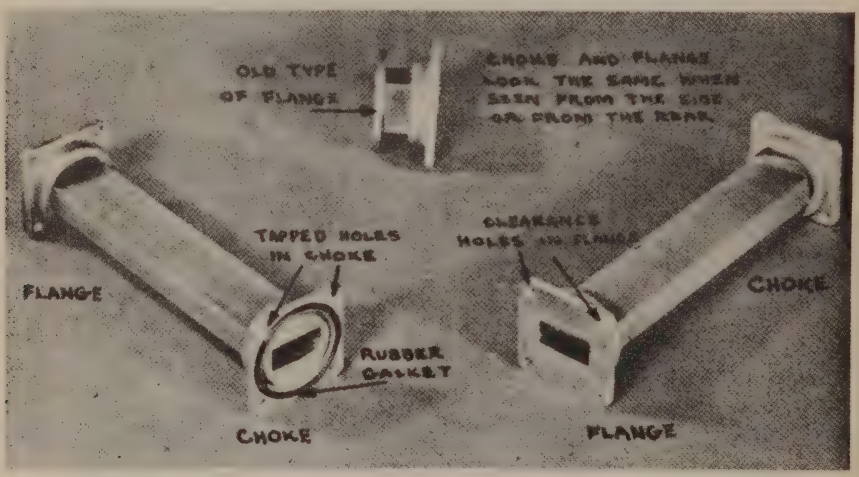


FIGURE 108.—Pressurized choke-flange joint in 1½-inch x 1-inch wave guide.

tems only the components most likely to break down are pressurized. If the line is of wave guide, the pressurized portion can be isolated by means of a thin sheet of mica or polystyrene (mica preferable) between a pressurized choke and flange joint. If the line is coaxial, a bead might be used as a pressure seal in the line, but a bead is likely to cause break-down, so that in this case it is better to transform to wave guide and make the seal in a choke flange joint, as described above.

7. SPINNER HOUSING DESIGN CONSIDERATIONS

To protect the spinner from rain and wind a radar housing, or *radome*, is used on almost all radar sets. The reasons for which a radome is used are of course purely mechanical in every case; but the electrical operation of the set is so dependent on housing design

that it must be regarded as an integral part of the r-f system, and not as a mere cover. Besides reducing the gain of the system by absorbing and reflecting some of the power, a radome always modifies the pattern to some extent, and it reflects enough power back into the line when poorly designed, to cause a serious frequency shift of the magnetron. If it were constant, this frequency shift, or *pulling*, could easily be compensated by an appropriate adjustment of the local oscillator; but in practice it is always variable, and constitutes an r-f problem of considerable importance.

To see the necessity for a constant frequency in the magnetron output, we note that the pass band of the i-f receiver is usually quite narrow (two megacycles is a typical width) while its input signal has a frequency equal to the difference between the frequency of the local oscillator and that of the transmitting magnetron. A change of a

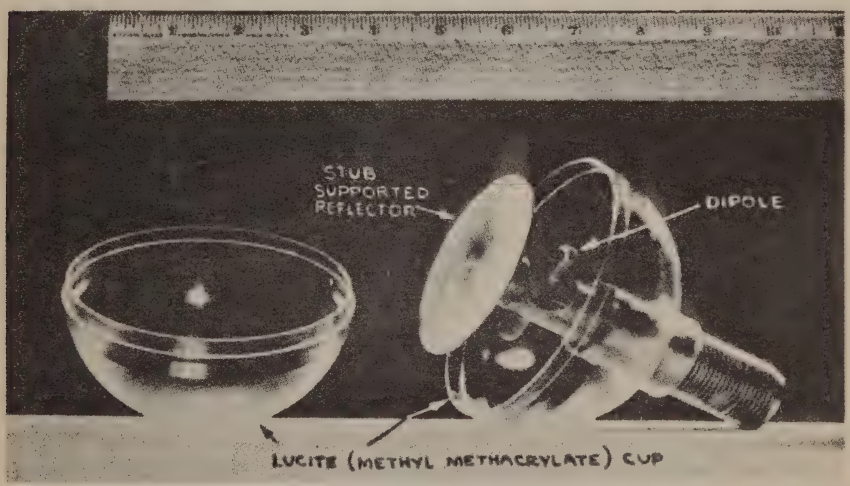


FIGURE 109.—Pressurized antenna on 7/8-inch coaxial line.

given magnitude in the magnetron frequency will therefore produce a change of the same magnitude (not of the same percent) in the input to the i-f receiver; and a shift of, say, three megacycles will evidently have very undesirable consequences even though a shift of this magnitude amounts to only 1/10 percent at S, or to an even smaller percent at a shorter wavelength. Many cases are recorded in which a signal is known to have been obliterated altogether by such a frequency shift, and the need for some method of correction is apparent.

Several methods are available. For example, one can reduce the selectivity of the i-f receiver, so that a small frequency shift will have no appreciable effect on signal strength. Or one can decouple the magnetron to some extent, thus reducing the amplitude of the reflected wave reaching it from the radome. Still another method is

deliberately to introduce a fixed mismatch in such a way that the variation produced by the radome occurs in a region in which the magnetron is relatively insensitive to changes of load impedance. These methods all have one feature in common: while they may reduce or even prevent the harm caused by pulling, they introduce new effects of their own which are always injurious to system performance. An increase in the pass band of the i-f receiver produces a corresponding increase in noise level, for example, while any decoupling of the magnetron, sufficient to prevent pulling, is accompanied by a serious loss in available power. The method of adjusting the initial mismatch, moreover, not only causes a loss of power by reflection, but it also produces a permanent standing wave which is particularly objectionable when the system is designed for high output.

To circumvent these difficulties, two main methods are used at the present time, viz., the installation of automatic frequency control (AFC) and the use of nonreflecting, or nearly non-reflecting, radomes. The first method does not attempt to prevent the frequency of the magnetron output from changing, but makes these changes harmless by producing a corresponding change in the frequency of the local oscillator; and thus the difference between these frequencies, which is also the frequency of the i-f input, is held constant. This method has the advantage that it compensates for frequency instability in general, not merely for that caused by the radome; and in many cases it will doubtless be found desirable even when the radome is designed to produce a negligible amount of pulling. There are, however, certain limitations on AFC; besides being reliable less than 100% of the time, it has the further disadvantage that it is at present unable to follow rapid changes that are likely to occur with conical or electrical scanning.

Proper design of the radome, on the other hand, almost always results in a radome which will not only produce a negligible amount of pulling, but will also have high transmission and a very small effect on the pattern. Rapidity of scan of course produces no especial complication, nor is the operation of a radome likely to be reliable less than 100% of the time (unless it is blown off—an unfortunate occurrence which has been known to happen in the past). At least four main methods are available for the design of non-reflecting radomes, which are not relevant, however, in a discussion of the r-f system; and it is believed that radomes can be designed on a production basis at each of the three bands, to produce a negligible amount of pulling. as well as a negligible change in the gain and pattern.

IV. TEST EQUIPMENT AND MEASUREMENTS

1. SLOTTED SECTIONS

a. Standing Wave Detection in General.

In the steady state line condition where a line of characteristic impedance Z_0 is connected between a load of impedance Z and a generator of impedance, Z_g we can postulate two waves traveling on the line in opposite directions, one away from the generator toward the load (the incident wave) and one reflected from the load (reflected wave) traveling toward the generator. The voltage and current in the line at any point are simply the vector sum of the incident and the reflected waves at that point. If in figure 110, A represents the value of the incident current or voltage wave and B the respective reflected wave, we can find a point on the line where the vector sum $A+B$ is a maximum and equal to $|A|+|B|$. Likewise we can find a point where the vector sum is a minimum and equal to $|A|-|B|$. It is useful to define the standing wave ratio (SWR or r) as

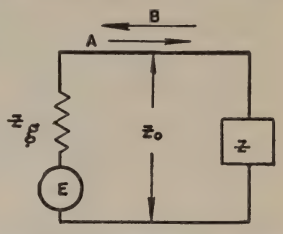


FIGURE 110.—Transmission line circuit.

$$SWR=r=\frac{\text{Max}}{\text{Min}}=\frac{|A|+|B|}{|A|-|B|}=\frac{1+\frac{|B|}{|A|}}{1-\frac{|B|}{|A|}}=\frac{1+|\Gamma|}{1-|\Gamma|}$$

From line theory we know that the reflection coefficient T is

$$\Gamma=\frac{B}{A}=\frac{Z-Z_0}{Z+Z_0}$$

and from this we may obtain the normalized impedance z.

$$z=\frac{Z}{Z_0}=\frac{1+\Gamma}{1-\Gamma}$$

By measuring r and noting the position of the minimum it is possible to measure the impedance of the load. Usually the reflected wave B represents a power loss, and in order to minimize this loss it is desirable to keep the standing wave ratio small in an r-f system. Curve 1 in the appendix shows the value of reflection loss as a function of the standing wave ratio. Furthermore, most transmitter tubes employed in the microwave region are very sensitive about the load into which they work. Changing the value of the load impedance into which such a tube is connected will cause it to change frequency and will often cause the output to vary. Since these tubes are engineered to connect to a matched load (one for which $Z=Z_0$) it is desirable to maintain such a load. All of the foregoing is to say that we need a gadget to measure standing waves.

In general the problem of measuring standing waves in radio frequency lines is very similar at all wavelengths from a few meters down to the microwaves, though the actual equipment used on the different wavelength bands will differ considerably in construction, refinement, operation and, naturally, size.

The elements of a typical standing wave detector set are shown in the block diagram figure 111.

The first unit (A) is the r-f power source which at microwaves is usually a velocity-modulation tube (see sections III, 3a and IV, 5) capable of being tuned over the band on which measurements are to be made. The oscillator is normally followed by an attenuator (B) to eliminate the pulling effect that a variable load would normally have on it. (C) is a slotted section in which a pick-up probe is moved

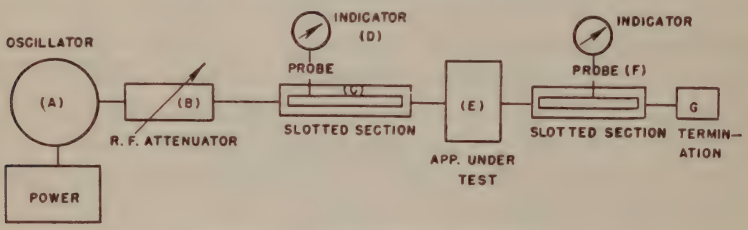


FIGURE 111.—Typical standing wave set-up.

along the line in the electric field. Mounted in the probe is a detector which supplies a d-c or audio-frequency potential to the indicator (D). The response of the detector-indicator combination is a measure of the field intensity at the probe; as the probe is moved along the line the meter gives an indication of the relative field strength of points in the section. If the equipment under observation is a transmission-type device, it will usually need to be terminated by a matched load (G). It is normal practice to place a slotted section and probe in front of the load so that the reflection from the load can be matched out as wavelength is changed. Usually a tuner is provided so that the impedance looking out of the second slotted section may be matched.

b. Coaxial Slotted Sections.

A recent design 10-cm. band coaxial slotted section for ⅞-inch line is shown in figure 112. The inner conductor of this unit is supported by a bead and by the inner conductor of the tuners which may be seen on the left-hand side of the slotted section. The other end, which is connected to the test piece, has no supports for they might cause reflections. Other slotted sections in common use employ stub supports (see section II, 2a) or polystyrene beads to support the center conductor. In a coaxial line when only the normal mode exists, the electric field lines are radial as shown in figure 113, and the probe, which projects into the field through a milled slot in the outer con-

ductor, is in a position to couple energy out of the line and into the detector.

The slot milled in the outer conductor is parallel to the axis of the line; and since the currents on the wall of the outer conductor travel parallel to the axis, the slot does not disturb the field inside the line. (This is true only if no higher modes exist in the line.) Also if the

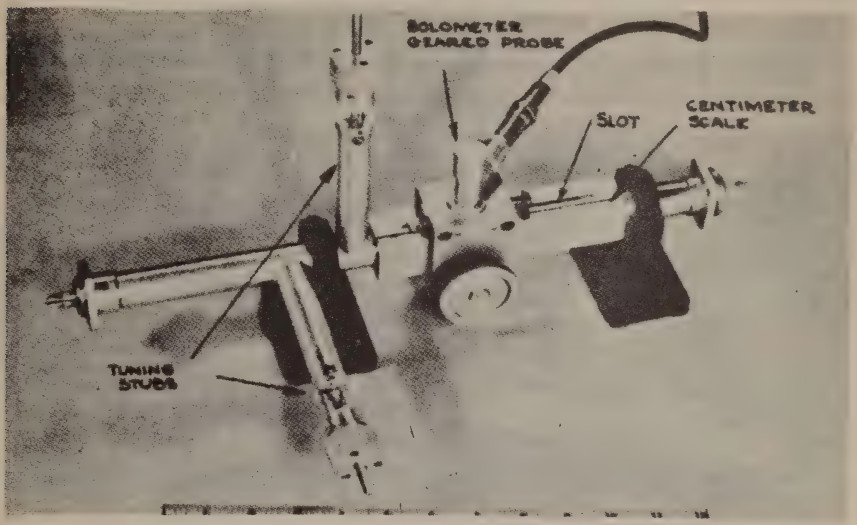


FIGURE 112.—Slotted section in ⅞-inch coaxial line.

probe is small and if it does not project into the slot very far its effect on the field is slight. On one section shown in figure 115, and, in fact, on most standing wave machines the probe is moved along the line by a rack and pinion attached to a knob which enables the operator to make fine adjustments of the probe position. Coaxial slotted sections similar to the ⅞–inch unit exist for 5/16–inch, ½–inch, ⅝–inch, and 1⅝–inch line.

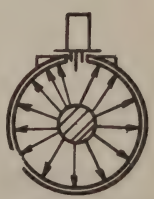


FIGURE 113.—Coaxial slotted section showing field probe.

Care should be taken during the construction of the equipment, and later when it is being used, to be sure that the inner conductor is well centered and that the probe projects into the slot the same distance over the entire length of travel. If these precautions are not taken, successive maxima and minima will have different values and an uncertainty will be introduced in computing standing wave ratios; that is, different ratios will be obtained by using different pairs of the maxima and minima.

The 5/16–inch and the ½–inch slotted sections can be used in the 3-cm. band, but the others (⅝–inch, ⅞–inch, and 1⅝–inch) are so large that they will support other than just the coaxial mode at 3–cm. and measurements made with them on this band would have little meaning.

Care must be taken when the 1⅝-inch equipment is used on the 10-cm. band, for higher modes can exist on this line at the short wavelength end of the band.

At 3 cm. wavelengths and below, coaxial lines, and therefore coaxial slotted sections, are used only for special applications, wave guide being the usual method of piping r-f about.

c. Wave Guide Slotted Sections.

Figures 114 and 115 show wave guide slotted sections used on the 10-cm. and 3-cm. band, respectively. In each case the section is

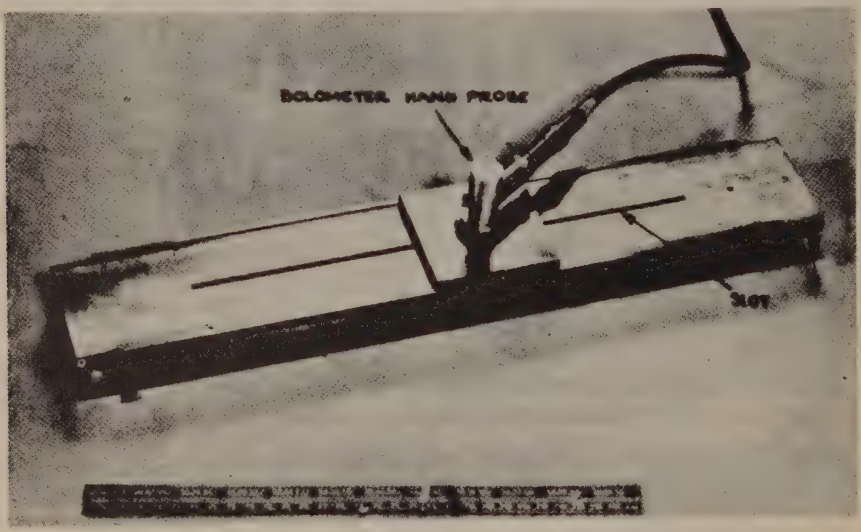


FIGURE 114.—Slotted section in 1½-inch x 3-inch wave guide.

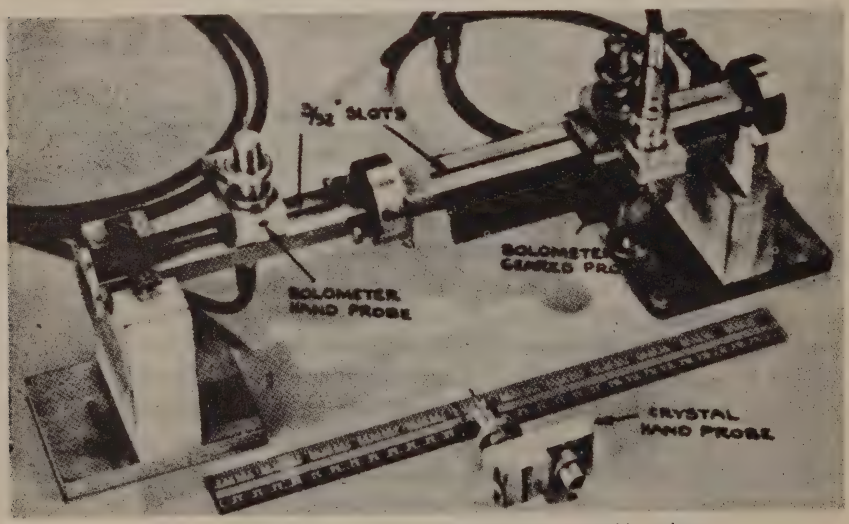


FIGURE 115.—3-cm. wave guide slotted sections and tunable probes.

designed for use with guide carrying the lowest ($TE_{1,0}$) mode. The current distribution in the walls of a guide transmitting energy in this mode is shown in figure 116. It can be seen from this figure that the currents flow in such a manner in the wave guide walls that a slot located at the center of the wide dimension as shown will not cause any of the current paths to be distorted, and so the fields within the pipe are just the same as they would have been had the slot not been made. If other modes were being propagated in the pipe this would not necessarily have been so, and a slot located at the center might

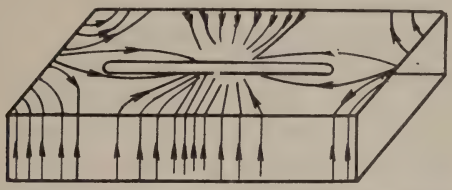


FIGURE 116.—Current flow in the wall of a wave guide propagating the lowest mode ($TE_{1,0}$).

cause serious problems. In addition, if more than one mode exists in the wave guide, interpretation of the results obtained by standing wave measurements is difficult; these problems are specialized, and beyond the scope of this book.

As in the case of coaxial slotted sections it is important that the surface which is slotted be uniform so that as the probe pick-up is moved along the slotted section the probe will always project into the guide the same distance.

Slotted section for round wave guide.—The slotted sections shown in figure 117 are intended for use with the $TE_{1,1}$ (H_1) mode. A larger

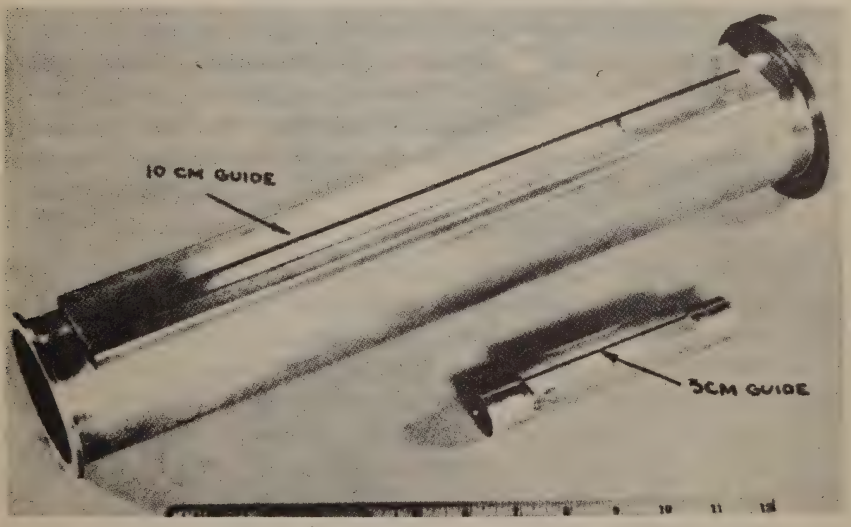


FIGURE 117.—Slotted sections in round wave guide.

size guide with a similar slot would be used for the $TE_{0,1}$ (E_0) mode when no other modes are present. The electric field distribution for these modes is shown in figure 118.

If the slotted section is to be used to measure standing waves in a $TM_{0,1}$ field the slot can lie anywhere on the circumference parallel to the main axis, but if the section is being used to measure $TE_{1,1}$ waves it is important that the slot be located at A or B [figure 118 (A)], that is at a point on the circumference where there are no current paths which are not parallel to the guide axis. A slot anywhere else (for example, at C) would disturb the normal field and yield results having little meaning. It is difficult if not impossible, to propagate this mode for any distance in round pipe without setting up a polarization normal to the main field. That is, we would establish an electric field whose polarization is along the CD line instead of along AB as shown above. The slot located at A would distort this field and set up reflections. When investigating such a field one should take care to be certain that the slot is at the proper location.

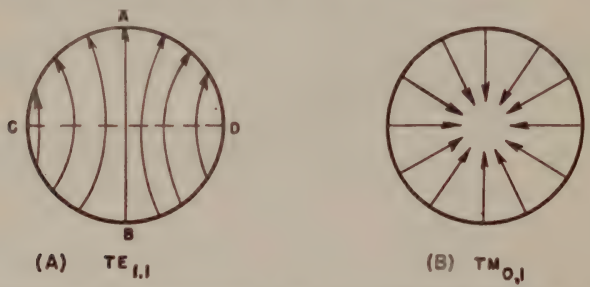


FIGURE 118.—Electric field distribution in the two lowest modes (round wave guide).

Higher modes are occasionally used in round pipes. Certain of these can be investigated by means of slotted sections if care is taken to get the slot at the proper point on the circumference. Other modes cannot be measured in this manner, and more elaborate methods, such as sliding sections with small probe holes, must be used to investigate the field patterns.

2. PROBES, PROBE DETECTORS, AND INDICATORS

Radio-frequency energy is normally extracted from the slotted section by the projecting probe, and usually transferred to a detector by means of a short length of coaxial line. A number of different elements have been used in standing wave detector probes to convert the radio-frequency energy to direct current or audio-frequency signals so that indication can be obtained on a meter. The most commonly used elements are crystals, thermocouples, and Littelfuse and thermistor bolometers. The crystal and the thermocouple are frequently used in conjunction with a sensitive microammeter to give indications which

do not depend on modulating equipment or amplifiers. Use of a Littelfuse generally requires that the signal source be modulated and that an amplifier precede the final rectifier which is usually a diode such as the 6H6.

Both types of bolometers have been used with balanced bridge circuits to measure standing waves, but this method of measurement is tedious and so not often used.

a. Crystal Probes.

A crystal connected as shown in figure 119 will rectify the radio-frequency energy transferred to it by the probe, producing a current in an external circuit that is approximately proportional to the square of the radio frequency field at the probe (proportional to the power at the probe).

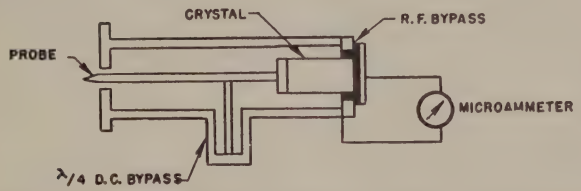


FIGURE 119.—Radio-frequency and direct-current circuits in a crystal probe.

A crystal detector of this type and a good low-resistance microammeter (Weston Model 440, 50 ohm, 0–30 microamperes) are often used for standing wave measurements in the 3 cm. band. On the 10 cm. band, where more power is available from the bench oscillator, higher resistance meters have been employed.

If the radio frequency source is modulated (usually 1,000 to 2,000 cycles/sec.), the crystal current will contain an audio-frequency component corresponding to the modulation frequency which can be amplified and then rectified; the output of the rectifier being placed on a meter. This system, while somewhat more complicated than the probe-microammeter combination, permits operation with smaller probe projections, allows smaller signals to be detected, and where the added complication can be permitted, is more satisfactory. If the amplifier and rectifier have a linear characteristic, the response of this combination of apparatus is also proportional to the square of the field at the probe.

It was stated previously that the crystal response is approximately square law. It is so close to square law in fact that if a probe employing a crystal is to measure the normally encountered standing wave (standing wave ratios of 4 or 5 in power—$r^2=4$ or 5) little error is ever encountered in making this square-law assumption. However, if it is desired to measure large standing waves or to make accurate measurements of transmission loss using a crystal, the crystal should be calibrated. A curve of crystal current as a function of

input power (see figure 120) is employed in such instances to convert the crystal current readings to actual power values. In figure 120 the maximum abscissa value is of the order of 10 microwatts.

Figure 121 shows crystal current as a function of larger powers than in figure 120.

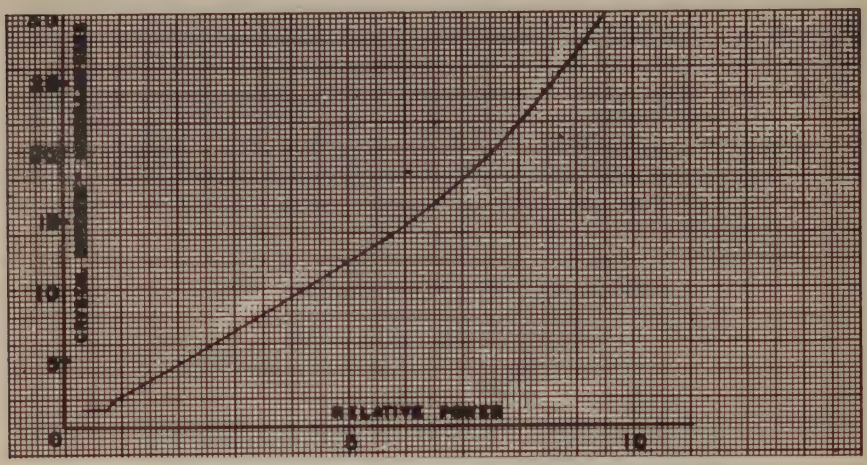


FIGURE 120.—Typical crystal calibration curve at low power.

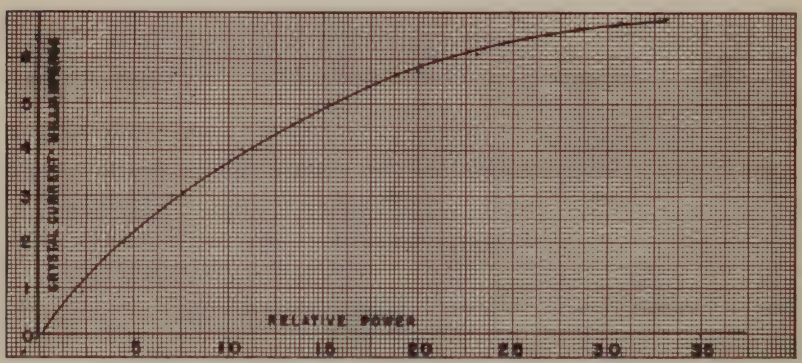


FIGURE 121.—Crystal current as a function of large powers.

Crystal calibration.—Three methods commonly employed to calibrate crystals will be described briefly. If a calibrated microwave attenuator is available, it is convenient to employ the arrangement pictured in figure 122.

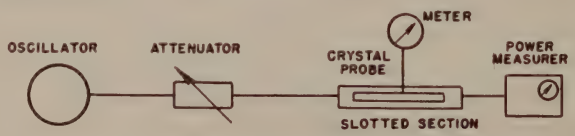


FIGURE 122.—Crystal calibration using calibrated attenuator.

With the attenuator set at the position of minimum attenuation the signal is adjusted by means of the first attenuator to a value producing the maximum crystal current to be encountered in operation. After the value of the crystal current is noted, the calibrated attenuator is used to reduce the power by small known increments. The various crystal current readings are plotted against the powers related to the maximum value. The curve obtained will be similar to that shown in figure 120.

If an accurate calibrated attenuator is not available, the crystal can be calibrated with the aid of a bolometer power measuring device. Figure 123 shows the arrangement that can be used for this measurement.

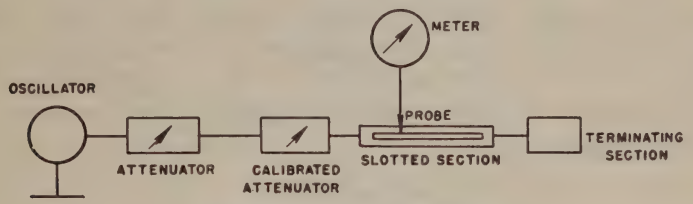


FIGURE 123.—Crystal calibration.

The terminating section used in the previous method is replaced by the bolometer; as the power into the system is varied the crystal current and the power in the line are recorded and the data plotted.

An especially simple and practical method of calibrating crystals uses only the equipment normally used for standing wave measurements (See figures 111, 192, and 193). A slotted section (standing wave machine) is first terminated in a short circuit at the output end. This causes an infinite standing wave ratio to exist in the line or guide. The line voltage then has zero amplitude at the short circuit and at the voltage minima which occur every half wave. length. The amplitude varies as the cosine of the electrical angle- $2\pi l/\lambda$, where l is measured from the position of a voltage maximum. A series of currents I are obtained (on the crystal-current microammeter or amplifier-output meter) a d the corresponding values of l noted. The values of l permit calculation of the voltage fed into the crystal (on a relative scale) and can be used for an I vs. V plot of the crystal characteristic curve. If plotted on log-log paper, the slope gives the so-called "law" of the crystal.

A convenient procedure is as follows: (1) Use the same crystal current meter or audio amplifier that will be used with the crystal in practice. Be sure the amplifier gain or meter is that normally used, as the law of a crystal depends on its power level and hence on the amplifier gain or meter sensitivity and resistance. (2) Adjust the power into the crystal (by varying probe insertion or line-power level) to give full scale at a voltage maximum. Be sure the probe

insertion is not enough to disturb the standing wave pattern as it is moved along; this may be tested by monitoring with a second probe. (3) Pick a convenient set of output current readings and determine the probe positions on either side of the chosen maximum corresponding to each current. The difference in probe-position scale readings on the two sides of the peak is equal to $2l$ for that current. (4) Use the relation $V \propto \cos 2\pi\, l/\lambda$ to calculate the relative voltages for each value of current. (5) Plot I vs. V. A log-log plot should be used if the crystal law is desired.

Several simplifying schemes may be used if one is content to take the law by calculating the slope of the log-log relation between two chosen points rather than plotting the whole I vs. V curve.

Another method, good for approximately square-law detectors, makes use of the linearity of an expansion of $V^2=\cos^2\theta$ in the neighborhood of half-of-maximum current ($\theta\pi/2$). One simply measures d_1, the separation of the two half-of-maximum current points on either side of maximum. Then

$$n\sim\frac{0.220}{\frac{d_1}{\lambda}-0.141}$$

For the range $1.8 < n < 2.4$, the error in n is below 0.02. For instance if $n=2$, $d_1=0.25\lambda$ and the formula gives

$$n=\frac{0.220}{0.250-0.140}=2.00.$$

In speaking of a crystal "law" one assumes the rectified current to be related to the impressed voltage by

$$I=CV^n$$

where C is a proportionality constant and the exponent n is referred to as the "law" of the crystal (e. g., $n=2$ is called a square law).

The slope (on a log-log plot) between two points (I_1, V_1) and (I_2, V_2) gives this exponent, thus:

$$\log I_1=\log C+n\log V_1$$
$$\log I_2=\log C+n\log V_2$$
$$\log I_1=\log I_2=n(\log V_1-\log V_2)$$

so

$$n=\frac{\log I_1-\log I_2}{\log V_1-\log V_2}=\text{log}-\text{log slope}$$

If I and relative power are plotted on a log-log scale, as would be natural in the first two methods, n is twice the log-log slope:

$$I=CV^n=C'P^{n/2}$$

so

$$n/2=\text{log-log slope on log } I \text{ vs. log } P \text{ plot}$$

Care must be exercised in using a calibrated crystal, for mechanical shock, electrical overloading, or other abuses can change the crystal's characteristic considerably and thus render the calibration curve useless.

Care must be taken when the crystals are used for measurements on pulsed power to see that the crystal is not overloaded. Figure 121 shows the response on a typical crystal to large r-f signals. The maximum value of the abscissa on this curve represented approximately 100 milliwatts of power; this saturation characteristic will vary, of course, from crystal to crystal. In pulsed power work where the average currents are small, the tendency is to allow large peak powers to be imposed on the crystal. It is apparent that increases in power beyond the amount necessary to saturate the crystal will not cause any increase in the output of the unit, and thus if the crystal is badly overloaded, it is possible to have bad standing waves in the line and yet to have the standing wave detector indicate that the load was well matched.

Crystal probes—10 cm. band.—For operation in the 10 cm. band crystals are normally placed into the 10 cm. band bolometer probe (to be described in detail in the section on Littelfuses) by means of a special metal adapter sleeve which is placed over the crystal pin. Figure 124 shows a crystal with this adapter mounted on it. The adapter makes the total size of the crystal the same as that of the Littelfuse. Probes using a crystal have considerably higher output than bolometer probes (normally 10 to 20 db), and so can be used with low-gain amplifiers. As indicated previously, the probe response will be only approximately square law, but for most routine measurements, testing, etc., this limitation will not matter. Probes of this type have little frequency sensitivity.

FIGURE 124.—Crystal with adapter for S-band bolometer probe.

Crystal probes—3 cm. band.—The crystal probe used for 3 cm. band measurements is shown in figure 125. This view shows the construction; an external view is given in figure 109. The probe is made in two styles, one to be used on geared traveling rack mounts, and the other to be used with simple slotted sections and held by hand.

As may be seen by reference to figure 125 the crystal (f) is mounted in a section of wave guide (a) approximately $\lambda/4$ from the end. Power from the wave guide or coaxial line being investigated is picked up by the probe (c) and transmitted to the wave guide (a). Adjusting the height of barrel (d) and plunger (e) allows one to adjust the probe so that maximum possible energy is transferred into the wave-guide chamber, coupling from the section

being tested is adjusted by controlling the distance which probe (c) projects into the guide.

The condenser formed by the 0.005-inch poly-tape at (g) and the r-f choke (b) provide an r-f bypass without also shorting the direct- and audio-frequency currents. All that was said in the previous paragraphs regarding crystal response applies to crystals used in these probes.

The tunable 3 cm.-band probe is quite frequency sensitive ($Q=300$), and needs adjustment when the operating frequency is changed.

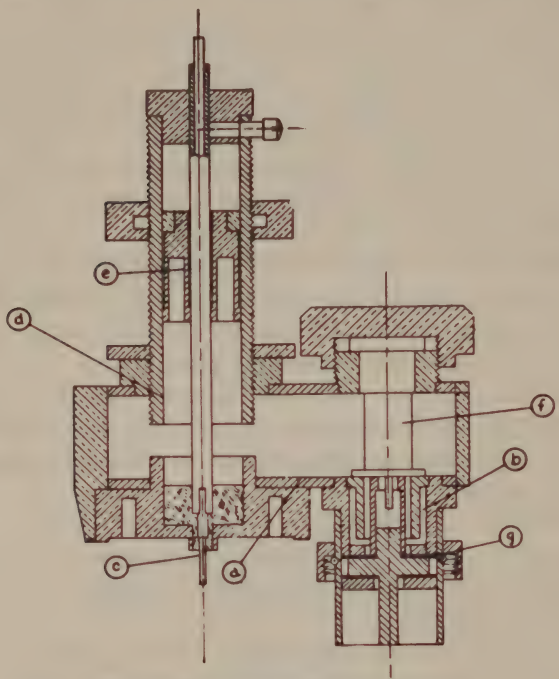


FIGURE 125.—Crystal probe for 3-cm. band.

b. Bolometer Probes.

"Bolometer" is a name given to a large class of gadgets used in radio-frequency measurements that have one property in common; namely, their resistance changes with changes in temperature; a change brought about in all cases with which we are concerned by the absorption of r-f power. Ordinarily the change is in a positive direction, that is, the resistance of the bolometer increases with temperature. However, there are certain materials which are used for bolometers which have negative coefficients. These are not generally used for standing wave measurements and will not be described here.

Nearly all conducting material have a resistance which is a function of temperature. It is this property of the platinum-wire bolometer which makes it useful for power measurements as well as for meas-

uring standing waves. Most common of all bolometers used for microwave measurements is the Littelfuse, a low-current, instrument protecting, fuse consisting of a very fine platinum wire mounted in a glass cartridge. Figure 126 shows the internal construction of the fuse. The upper cap and protecting glass envelope have been removed to show the insides. The two heavy leads held together by the glass bead are supports; the platinum wire is soldered between them as shown.

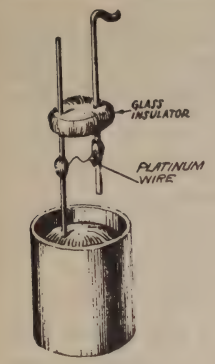


FIGURE 126.—Littelfuse without upper cap and glass (about twice size).

$\frac{1}{32}$ amp., $\frac{1}{100}$ amp., and $\frac{1}{200}$ amp. Littelfuses are commonly used for bolometers. Figure 127 shows the way the resistance of typical fuse elements varies with applied power. To vary the resistance it is necessary only that heat energy be supplied to the element. The heat can be supplied by d-c, radio-frequency signals, or other electrical currents. The time constant of these units is very small (approximately 1/20,000 sec. for the 1/200 amp. fuse), and so if the heating power is modulated at a low audio-frequency rate (1000–2000 cycles/sec.), the resistance of the fuse will be modulated at that rate. For example, if the fuse was being heated by microwave energy mod-

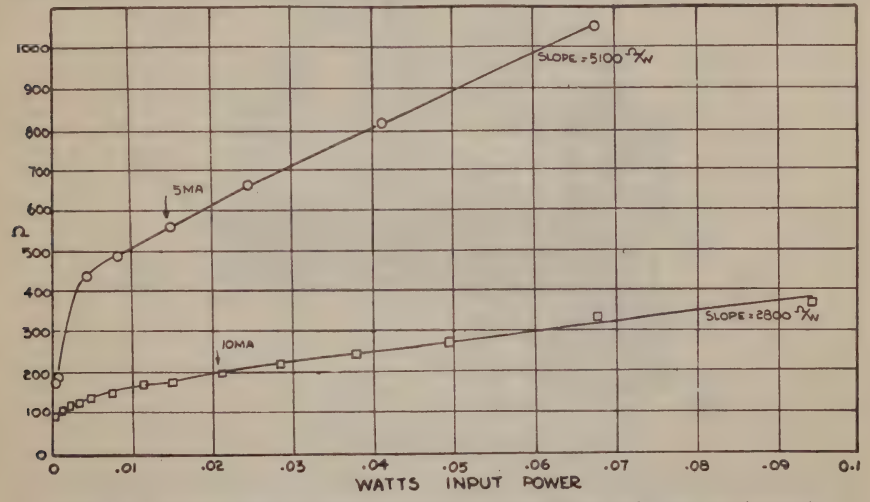


FIGURE 127.—Direct-current calibration curves for 5 ma. fuse (circles) and 10 ma. fuse (squares).

ulated at 1000 cycles per second the resistance of the fuse would vary in the same manner as the radio-frequency envelope that is, at 1000 cycles/sec. The amount of resistance change will be a function of the slope of the resistance-power characteristic (the rate at which resistance changes with applied power—measured in ohms/watt), the amount of applied power, and the percentage modulation.

The changing resistance is normally employed to modulate a small

direct current and thus provide an audio signal that can be amplified. A schematic of the circuit usually employed to accomplish this is shown in figure 128.

The d-c power source V causes a small current to flow through the Littelfuse (r_1) and the transformer primary (r_2); this current is modulated when the fuse resistance is changed by the superimposed radio-

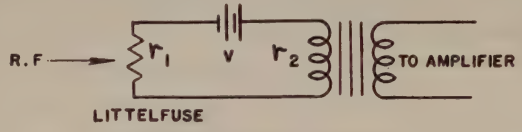


FIGURE 128.—Littelfuse bias circuit.

frequency power; and this varying current flowing in the primary of the transformer causes an a-c voltage to be produced on the secondary side of the transformer, which in turn is used to actuate a vacuum tube amplifier and rectifier.

The response of the bolometer is accurately square law, and so there is generally no need to calibrate them.

Bolometer probes—10 cm. band.—The constructional details of the bolometer probe commonly used on the 10 cm. band is shown in

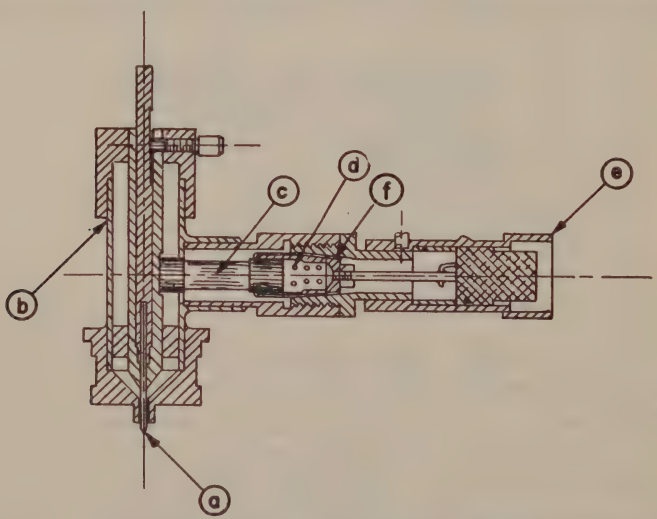


FIGURE 129.—10 cm. band bolometer probe—fixed tuned

figure 129. It consists of a probe pick-up (a) which projects into the slotted section, a section of coaxial line (b) approximately ½λ long, a Littelfuse bolometer (c), an r-f bypass (d), and an Amphenol cable connector (e). The bolometer makes contact with the ½λ line at a point approximately ¼λ from the top end of the line, that is, at a voltage maximum. The other end of the fuse fits into a clip holder

which is insulated by the poly-tape sleeve (f) to prevent a d-c short circuit. Since the capacity is fairly large at this point it provides an adequate r-f bypass. The audio and d-c circuits are completed to the amplifier by means of a cable connected at (e). This probe is not tunable, is low Q, and therefore has little frequency sensitivity, its response being adequate over the 7–14 centimeter region. A slight variation of this design provides the probe with a plunger at the top end that can be adjusted, this provides an adjustment to match the fuse impedance at specific wavelengths. With such a unit it is possible to get greater output, and thus be able to make measurements with less sensitive amplifiers than is possible with the untuned probe.

Bolometer probes—3 cm. band.—Figure 130 shows a 3 cm. band

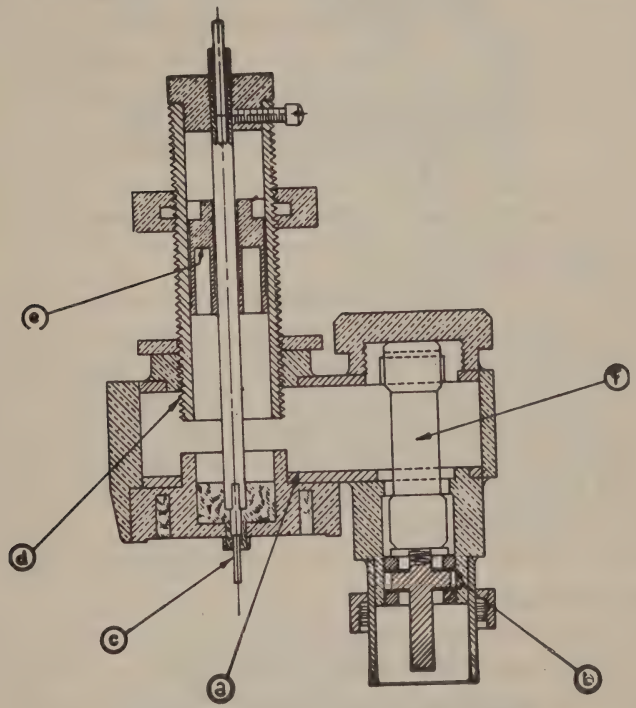


FIGURE 130.—3-cm. band bolometer probe—tunable.

bolometer probe that is normally used only for high level, pulsed power measurements, though it is adaptable to low level work if sufficient r-f energy or sensitive enough amplifiers are available. In Figure 130 the fuse (f) is mounted in the wave guide as shown, (b) are r-f bypass condensers, probe (c) feeds the coaxial tuning section composed of barrel (d), and plunger (e). By adjusting the positions of the plunger and of the barrel with respect to the bottom of the guide the bolometer is patched to the probe. This unit has a Q of approximately 300 and so must be readjusted each time the frequency of the power source is

changed. Figure 131 shows the construction of an untuned 3 cm. bolometer probe. The coaxial tuner in the previous probe has been replaced by the antenna (a) which does not make adjustments for the bolometer mismatch and is therefore less sensitive than previously (this probe is about 10 db less sensitive than the previous one). Its response changes only about 3 db over the band 3.16 to 3.24 cm.

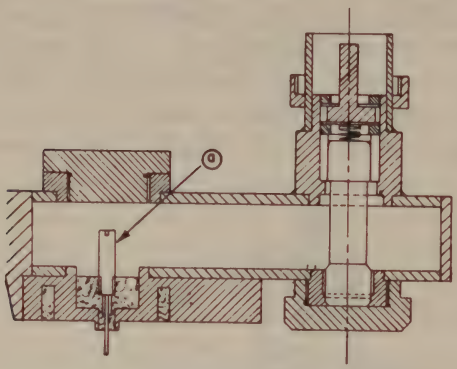


FIGURE 131.—3-cm. band bolometer probe—fixed tuned.

c. Thermocouple Probes.

Historically the thermocouple probe is important for it provided the first means of measuring standing waves. Such a device usually consisted of a very sensitive vacuum thermocouple mounted in a small cavity and fed by a probe in much the same way as the probes previously described. However, the thermocouple junctions were very hard to match to the probe, they were delicate mechanically, and they were quite insensitive and sluggish because of their high thermal capacity. In addition to all of this they were very expensive and so have almost entirely given way to the previously described elements.

d. Meters and Amplifiers.

Meters.—Crystal and thermocouple probes have often been used with sensitive microammeters to give amplitude indications. This is a satisfactory arrangement when sufficient power is available from the signal generator to provide usable meter deflections without requiring probe lengths sufficient to disturb the measuring system. This condition can usually be met at 10 centimeters using a tunable crystal probe and a 417 or 707A power source, and at 3 centimeters using the tunable crystal probe and an r-f source such as the 419 Klystron or the British KRN–2, or the Bell Telephone Laboratories 1280–CT (Samuel's tube). Using a 723A/B tube at 3 cm. wavelengths is not always safe, for the 723A/B does not put out enough power to allow sufficient isolation in the form of attenuation to be inserted between the oscillator and the component being tested. Insufficient isolation may result

in "pulling" of the generator as the component is adjusted. This probe-and-microammeter technique is often employed for field testing where amplifiers are undesirable. With care accurate results may be obtained by this method of measurement.

Amplifiers.—For use as indicators with standing wave machines there are four main requirements which an amplifier must fulfill:

(*1*) *Sufficient gain.*—For use with a Littelfuse bolometer probe on a low-level system, an amplifier must have a sensitivity of 10 microvolts full-scale or better. A crystal probe for 10 cm. requires a sensitivity of approximately 200–400 microvolts.

For use with a 3 cm. crystal probe a sensitivity of 40 μv. is desirable if the amplifier is used with a 723A/B local oscillator. If a 419 is used as the power source, 100–200 μv. sensitivity is adequate. These figures assume that the r-f source is modulated heavily; i. e. 50 percent modulation or better.

For measurements on high-level pulsed power using a bolometer probe a sensitivity of 5 millivolts is usually adequate, while for measuring low-level pulsed power (20 milliwatt peak power or less) sensitivities of 1 microvolt or greater might be needed.

(*2*) *Linearity.*—The readings on the output meter should be proportional to the input voltage, deviations from a linear characteristic can introduce large errors in computing standing waves. If it is necessary to do work with a nonlinear amplifier, a calibration should be made of it.

(*3*) *Low noise and hum pick-up.*—The tube noise and extraneous audio pick-up when probes are connected and all of the associated apparatus is running should be so small that with the amplifier adjusted for maximum sensitivity the noise causes less than a 1-percent meter deflection. If it is necessary to use an amplifier with a small amount of output noise, the meter point should be reset to zero if possible. The amount of noise registered will usually be a function of gain setting.

(*4*) *Stable gain.*—For accurate measurement work the amplifier gain should be independent of line voltage changes.

K-1 amplifiers.—Figure 132 shows the circuit diagram of the K-1 amplifier used at the Radiation Laboratory. The amplifier employs two pentode amplification stages followed by a 6H6 rectifier driving a 0-1 ma. meter.

The input circuit is arranged so that either of two probes may be connected to the input (switch S_2). Individual gain controls are available for each probe. Power from the main supply is brought through resistor R_4, to switch S_2 where it can be applied to the probe circuit to provide bias if a bolometer probe is connected to the amplifier.

The condenser-resistor network between the plate of the 6SJ7–GT

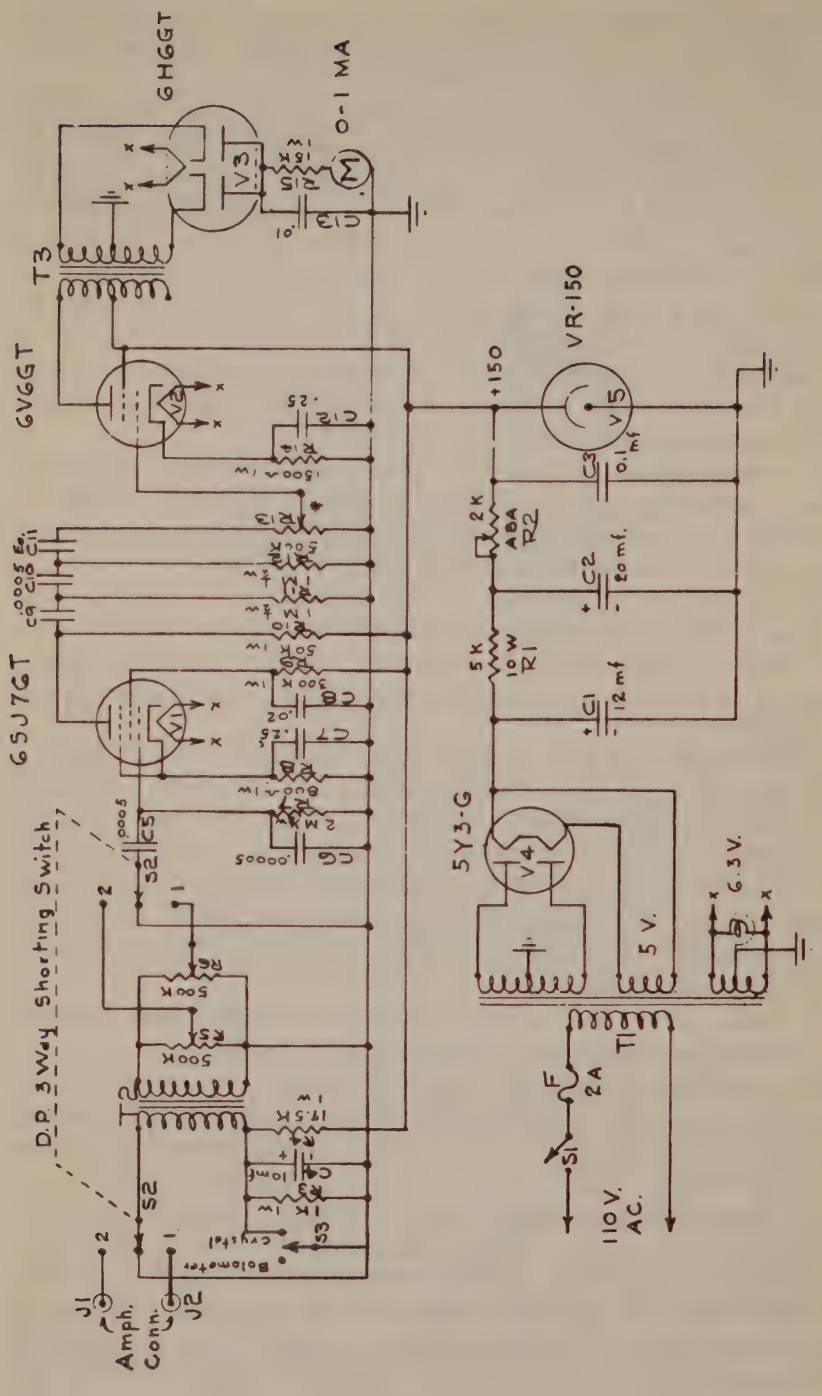


FIGURE 132.—Type K-1 amplifier circuit.

and the grid of the 6V6 reduces the low frequency gain of the amplifier, and thus reduces the noise and hum output of the amplifier. The amplifier has a full-scale sensitivity of approximately 5 millivolts and is linear. Some models of this unit have a meter dial calibrated to read voltage standing wave ratios.

Type "O" amplifiers.—A higher gain amplifier employing three stages of pentode amplification is shown in figure 133. The last stage is followed by a 6H6 rectifier and a 0–1 milliampere meter. The input circuit is similar to the input on the K–1 amplifier. To eliminate noise and hum a twin T inverse feedback circuit is employed from the plate to the grid of the second stage tube. At the resonance frequency of the network no signal is fed back to the grid but at all other frequencies a degenerative signal is applied to the grid reducing materially the gain of the amplifier. This produces an amplifier with a narrow band pass eliminating much noise, extraneous pick-up, etc.

The VR–150 tube provides regulation of the B+ supply. This amplifier has a full-scale sensitivity of 20–40 microvolts.

Type "P" amplifiers.—The twin T inverse feedback circuit of the type O amplifier limits its usefulness to a narrow band of frequencies. The type P amplifier was designed to have a frequency response which is practically uniform from 400 to 4,000 cycles per second so that it can be used with almost any system. It uses three stages of amplification followed by a 6H6 rectifier and meter circuit. See figure 134. The full-scale sensitivity is about 7 microvolts. The amplifier is linear and the scale is calibrated so that standing wave ratios in voltage can be read directly.

It has two inputs which can be fed by either crystal or thermistor type probes. Each input has a continuously variable gain control, and there is also a continuously variable master gain control. In addition there is a three-step gain control which can cut down the sensitivity by a factor of 100.

Troublesome ground currents have been avoided in the type P amplifier by tying all of the critical grounds to one common point and tying this point to the chassis through a 5-ohm resistor. The Amphenol input sockets are insulated from the chassis and grounded through this common point. In order to ground the outer conductor of the input cable with respect to video currents the Ampheno socket is connected to the chassis by means of a bypass condenser.

A variation of the type P amplifier is now being developed which promises to give a full-scale sensitivity of the order of 2 microvolts. This amplifier employs a variable inverse twin T feedback circuit (as in the type O amplifier) which can be tuned over the 400 to 4,000 c. p. s. range.

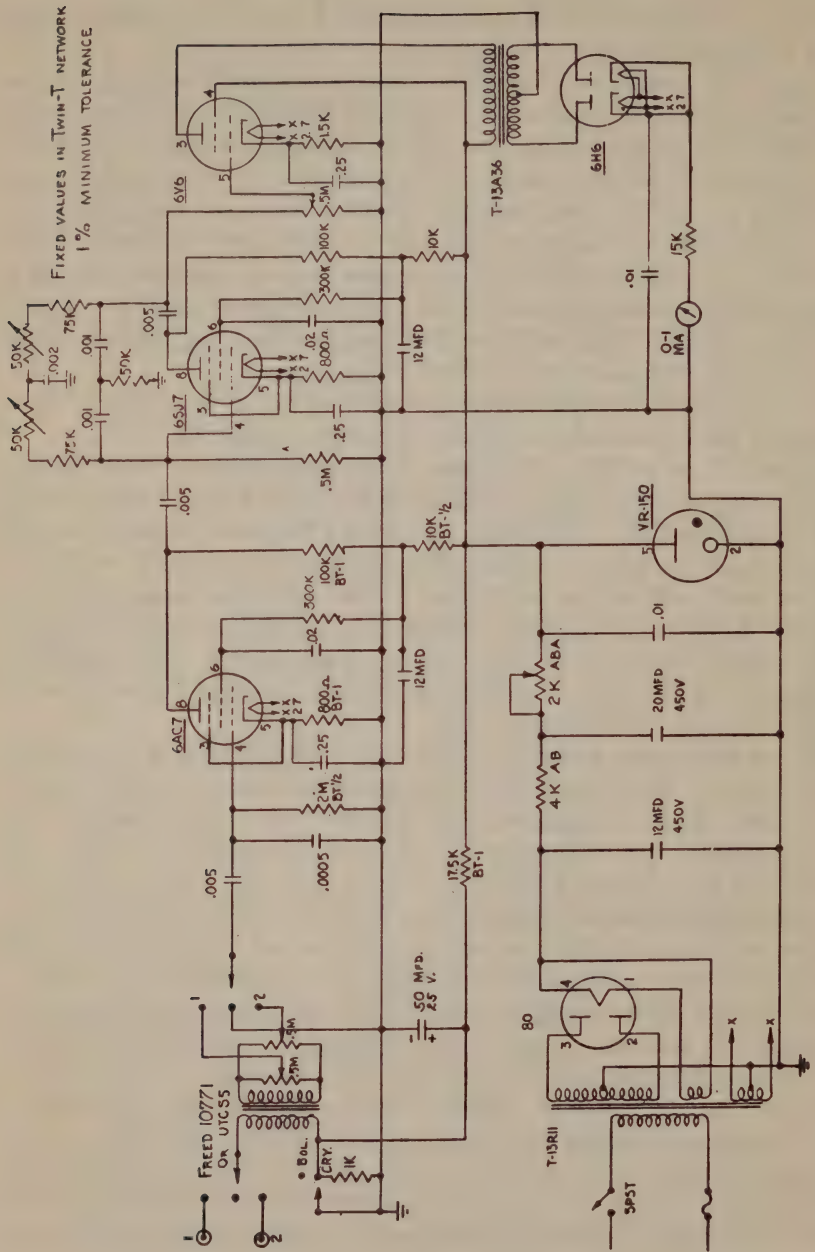


FIGURE 133.—Type O amplifier.

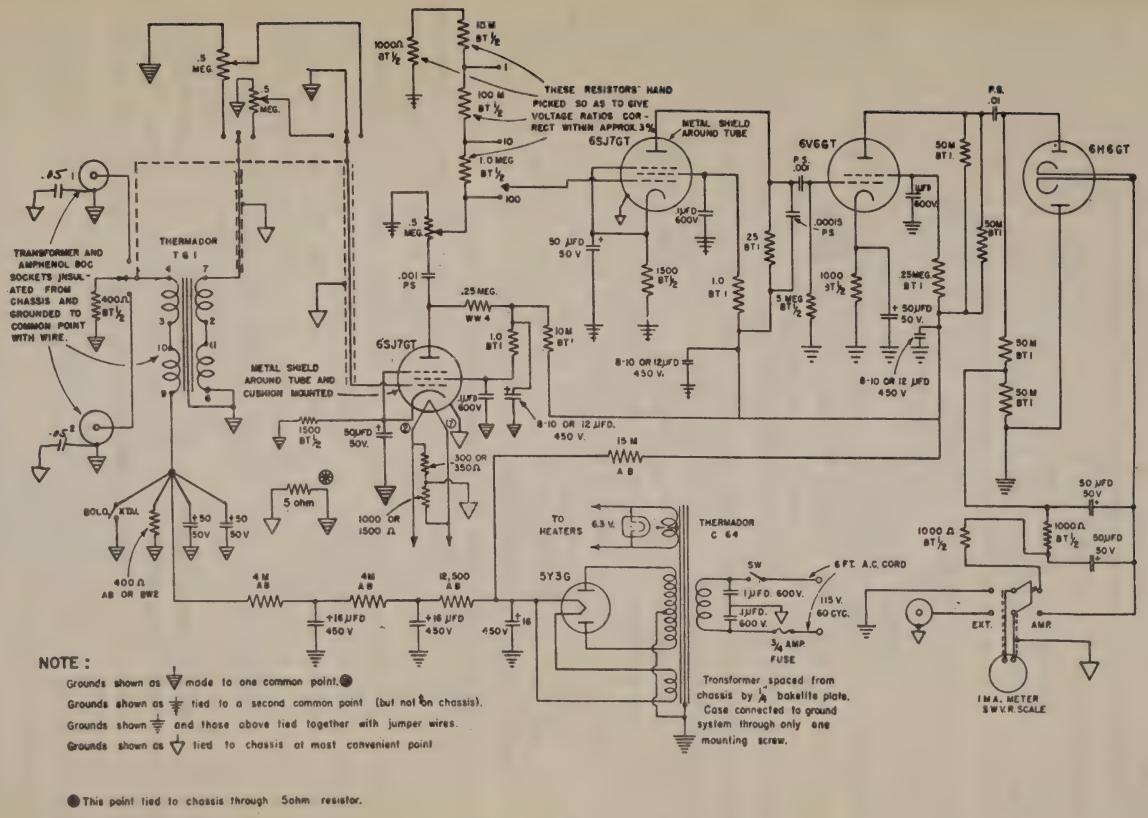

.5 MEG.
5 MEG.
.05 µF
TRANSFORMER AND AMPHENOL 80C SOCKETS INSULATED FROM CHASSIS AND GROUNDED TO COMMON POINT WITH WIRE.
THERMADOR T G I
400Ω BT½
.05 µF
METAL SHIELD AROUND TUBE AND CUSHION MOUNTED
1000Ω BT½
15 M BT½
100 M BT½
10
1.0 MEG BT½
100
THESE RESISTORS HAND PICKED SO AS TO GIVE VOLTAGE RATIOS CORRECT WITHIN APPROX. 3%
METAL SHIELD AROUND TUBE
6SJ7GT
6V6GT
6H6GT
.5 MEG.
.001 PS
.25 MEG. WW 4
1.0 BT I
10 M BT I
50 µFD 50 V
1500 BT½
25 BT I
.0001 PS
1 µFD 600V
1000 BT½
.25 MEG. BT I
50 M BT I
+ 50 µFD 50 V.
8-10 OR 12 µFD 450 V.
1500 BT½
50 µFD 50V.
300 OR 350Ω
15 M A B
5 ohm
1000 OR 1500 Ω
BOLO. XTAL
400 Ω AB OR BWE
+50 50V
+50 50V
4 M AB
4 M AB
12,500 AB
+16 µFD 450 V
+16 µFD 450 V
+16 450 V
TO HEATERS 6.3 V.
THERMADOR C 64
5Y3G
SW
6 FT. A.C. CORD
1 µFD. 600 V.
.1 µFD. 600 V.
115 V. 60 CYC.
¾ AMP FUSE
Transformer spaced from chassis by ⅛" bakelite plate. Case connected to ground system through only one mounting screw.
50 µFD 50 V
1000 Ω BT½
50 µFD 50 V
EXT.
AMP
I MA. METER S W V R SCALE
NOTE :
Grounds shown as ⏚ made to one common point.●
Grounds shown as ⏚ tied to a second common point (but not to chassis).
Grounds shown ⏚ and those above tied together with jumper wires.
Grounds shown as ▽ tied to chassis at most convenient point
● This point tied to chassis through 5ohm resistor.

FIGURE 134.—Type P amplifier.

3. VARIABLE IMPEDANCE MATCHING TRANSFORMERS

a. Coaxial Line Tuners.

Stub tuners.—To understand these devices one must know the following facts: First, a side branching coaxial line is in parallel with the main line since at the junction the voltage across the lines is common to the two branches and the current is shared between them. Hence the admittance of the stub adds to that in the main line. Second, the admittance of a shorted line (e. g. a stub) is a pure susceptance, varying from infinite susceptance at zero length through inductive values to zero susceptance at a quarter wavelength, then through capacitive values to infinite susceptance again at a half wavelength; hence a stub may add any desired amount of susceptance, either inductive or capacitive. Third, an understanding of some form of "admittance diagram" (same as used for impedances) is of invaluable assistance in tracing through the steps of admittance transformation at the stubs and in the lengths of line between stubs.

Various arrangements (numbers, spacing, etc.) of stubs have been proposed. The one in most general use consists of two stubs spaced three-eighths of a wavelength apart. A spacing of one-eighth wave length would probably be less frequency sensitive but is difficult to build, so the three-eighths wavelength spacing has become standard. The shorting plungers in the stubs have choke-type shorting plungers which employ contacts behind the chokes. The current which must be carried by the contacts is reduced by the presence of the chokes and consequently, trouble with contact fingers is greatly reduced. In adjusting such a tuner both stubs are first set at approximately the null (zero susceptance or quarter-wave) position. Then one stub is adjusted for best match (as indicated either by lowest standing wave ratio on a travelling probe between the tuner and the source of power, or by increased output, as the case may be). Then the second stub is used to improve the match further. A slight trimming of the first may then be needed, and so on until the desired degree of matching is achieved. If the standing wave ratio was initially less than two, such a tuner can always match the load to the line. If greater than two, the phase of the mismatch must be considered. With random phase about five out of six cases may be tuned even though the standing wave be very large. In general, we are dealing with less than two to one mismatches and so do not worry about this limitation.

The circular admittance chart (see IV, 10d and V, 3) is well suited to consideration of stub tuners.

Working out from matched input end in the schematic of figure 135:

$Y_0 = Y_1' + jB_2$ or $Y_1' = Y_0 - jB_2$

$Y_1 = Y_1'$ transformed through $\frac{3}{8}\ \lambda$ of line

$Y_1 = Y + jB_1$ or $Y = Y_1 = jB_1$

If Y_1' lies anywhere on the circle shown in figure 136 (e. g. at Q'), B_2 can be chosen so as to bring Y_1' to Y_0. The $\frac{3}{8}\ \lambda$ transformation toward the load is easy to see on this type chart; each point is merely rotated counterclockwise $\frac{3}{4}$ turn about the center, (e. g. $Q' \rightarrow Q$). (This rotation is to be interpreted as Q' transforms to Q.) Hence the Y' circle remains a circle, merely rotating $\frac{3}{4}$ turn. Rotations in the opposite sense represent transformations toward the generator, and the angle is always on the basis of one complete rotation per half wavelength. Now any admittance not lying in the "untunable" circle may be transformed (by adding the proper positive or negative susceptance B_1) so as to lie somewhere on the circle Y_1 (e. g. $P \rightarrow Q$). The untunable region is bounded by the circle $g=2$, which is the largest value of g for which a "g=constant" circle intersects the Y_1 circle. No admittances for which $r \leq 2$ lies within the "untunable" circle. Hence if $r<2$ we are certainly able to tune it out. Even for $r>2$ we can tune it out except where the phase is such as to cause the admittance at the stub 2 to lie within the "untunable" circle. Double-stub tuners are shown in figures 112 and 137.

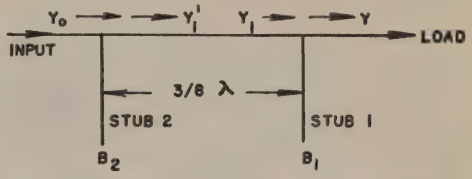


FIGURE 135.—Schematic drawing of a double stub tuner

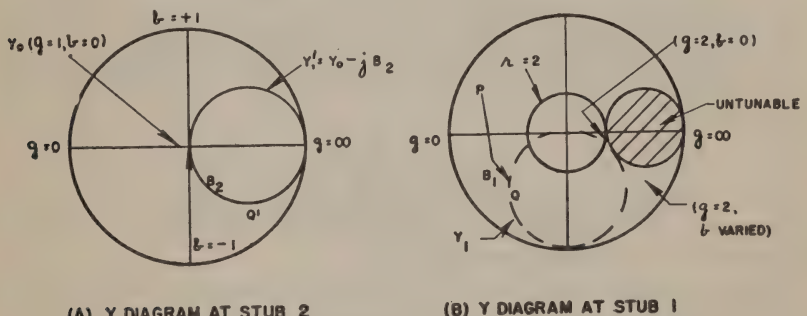


FIGURE 136.—Action of a tuner shown on an admittance diagram.

"Slug" tuners.—The simplest form of slug is a piece of dielectric material filling the space between the conductors and having a length equal to a quarter wavelength in the material. The characteristic impedance in the slug-filled line (call it Z_S) will then be lower than Z_0 by the factor $\sqrt{K}$. If placed in a matched line, this slug would transform Z_0 through a quarter wave of $Z_S = Z_0/\sqrt{K}$ to an impedance Z_0/K, which would give rise to a standing wave $r=K$. If such a slug were introduced into a line in which a standing wave $r=K$ already existed, a position could be found such that it would exactly compensate for the mismatch. In this position its load end would be at a voltage

minimum, where Z is also minimum and equal to Z_0/K. (See figure 138.)

This is precisely the situation if we have two such slugs separated by zero or an integral number of half wavelengths. In this case they act as a 1 : 1 transformer, as does any inserted section of line which is exactly half a wavelength long.

However, as the slugs are moved apart, they begin to introduce standing waves. The maximum is, as one expects, introduced when the spacing is a quarter wavelength. In this case, one slug introduces $r=K$, and the second further increases this to $r=K^2$. The impedance transforms as shown in figure 139.

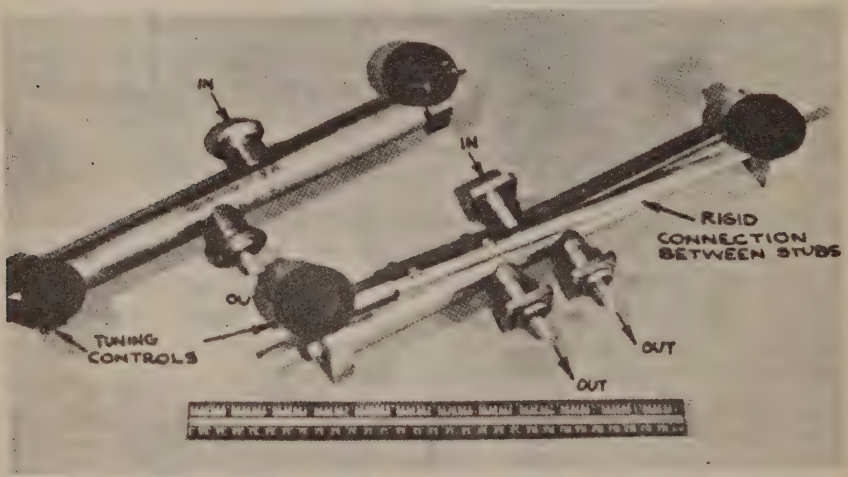


FIGURE 137.—Examples of the use of side stubs in coaxial lines. Left, double-stub tuner. Right, power divider.

The combination could, of course, be used to match out a standing wave $r=K^2$, by properly positioning it. The impedance would be as above with direction of power reversed.

Thus we see that such a pair of slugs may be made to cause standing waves from $r=1$ to $r=K^2$ by varying the spacing. And at each spacing, the combination may be used to match out a standing wave equal to that which it is capable of producing. To do this, it is only necessary to vary the position of the properly spaced slugs until the proper phase relations exist. Thus, this type of slug tuner may tune

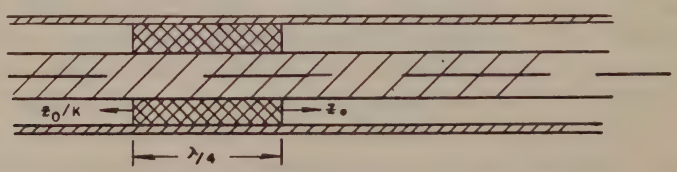


FIGURE 138.—Dielectric slug in a coaxial line.

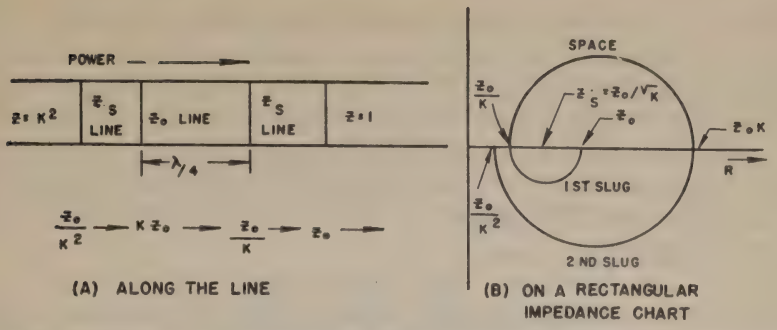
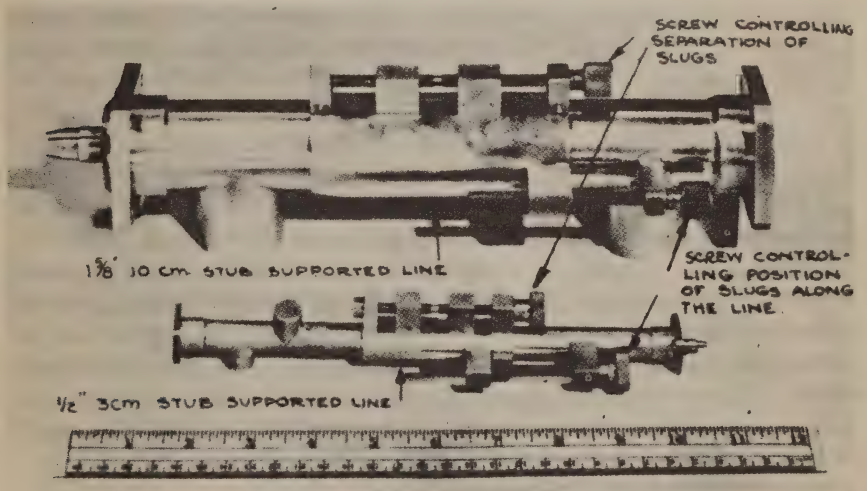


FIGURE 139.—Impedance along a double-slug tuner with a quarter wavelength spacing between the slugs.

out any standing wave from 1 to K^2. For polystyrene slugs this means any $r < K^2 = (2.6)^2 = 6.76$ may be matched.

In practice the slugs are moved by external screws connected to the slugs through a narrow slot along the length of the line. Two adjustments are provided: One for controlling the slug separation, the other for moving the two as a unit. These are adjusted alternately, each time for best match, until the desired degree of matching is achieved. Figure 140 shows two coaxial-slug tuners.


SCREW CONTROLLING SEPARATION OF SLUGS
1⅝" 10 cm STUB SUPPORTED LINE
SCREW CONTROLLING POSITION OF SLUGS ALONG THE LINE
½" 3cm STUB SUPPORTED LINE

If one desires a different maximum tunable r, different dielectrics or dimensions of slug may be used. Or, one may use other types of slugs. Metal slugs are of course less likely to be injured by sparkover, and can be made to give any desired tuning range. Contact trouble can be avoided by making the metal slugs in the form of sleeves which can be suspended between and coaxial with the two conductors. One such arrangement is shown in figure 141.

There is still the possibility of dielectric spark-over trouble. This may be eliminated (but at the cost of possible contact trouble) by fitting the slug against the outer conductor. Several slotted fingers at each end of the sleeve assist in making contacts. An all-metal slug for high power may be made by substituting capacity coupling for these contacts. In order to have room for the necessary chokes, the metal slug must be lengthened to ¾ λ. Such a design might be as shown in figure 142.

In any case, the tuning range is determined by the ratio of characteristic impedance in slug-filled line to that in the normal line. If we let this be $Z_S/Z_0 = 1/n$, then any $r < n^4$ can be tuned. For dielectric slugs $n = \sqrt{K}$, whence $r < K^2$, as shown above. For metal slugs $Z_S = Z_0 - Z_m$. Z_m is the "characteristic impedance" of the metal slug (i. e., 138 $\log_{10} r_2/r_1$ where r_2 is the outside and r_1 the inside radius of the slug).

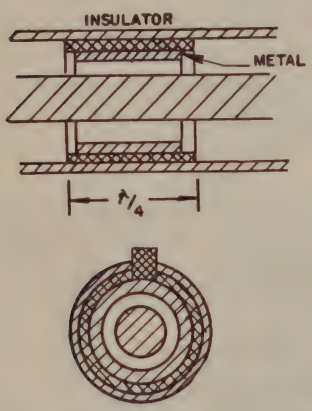


FIGURE 141.—One type of suspension for metal slugs.

A slug tuner has the advantage of being able to match any standing wave ratio up to any desired limit by using proper Z_S in its design. However, the larger Z_S, the more critical will be the adjustment, even for small standing waves. Also as Z_S increases so do the maximum fields; hence the transmittable power is limited. If the slugs are not identical it is impossible to avoid introducing a mismatch.

In certain cases (for instance in investigating the effect of load impedance on an oscillator) it is desirable to be able to introduce an arbitrary standing wave ratio and vary its phase. By choosing several values of standing wave ratio and varying the phase of each, the whole impedance plane may be conveniently mapped out. The double slug tuner is particularly well suited for accomplishing this. Starting with a matched load, the desired standing wave is introduced by properly spacing the slugs; then the phase is varied by moving the two slugs as a unit.

b. Wave Guide Tuners.

Stub tuners.—Just as in the case of coaxial lines a branching line may be used as a tuning stub. If the side stub branches from the narrow side of a rectangular guide the voltage (from top to bottom of the guide) at the junction is common to the two branches; and the current (that along the direction of propagation, which occurs along broad surfaces of the guide) is shared. This kind of branching obviously is of the shunt type, just as was the coaxial stub branch. Hence the discussion given there carries over exactly to this case.

On the other hand, if the stub line branches from the broad face of the rectangular guide the current along the broad face is interrupted and flows into the branch circuit thence into the output part of the main line in the manner characteristic of series circuits. It can also be seen that the voltage across the input guide equals the sum of the

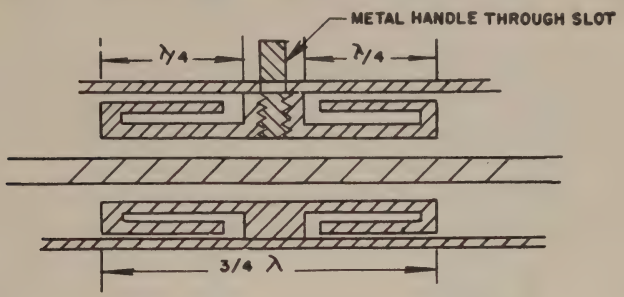


FIGURE 142.—Choke type of suspension for metal slugs.

voltages in branch and output part of the main line. In this series case, impedances add in the same way that admittances did in the shunt case. So if we replace admittance, conductance, and susceptance by impedance, resistance, and reactance the previous discussion applies to this series case.

In both types of branching, ⅜ λ spacing is a convenient one; this spacing makes certain the tunability of any $r<2$, as pointed out in section IV 3a.

Slug tuners.—Just as in the case of coaxial lines, slugs of various sorts may be introduced into the wave guide to give an impedance matching device. The calculations of tuning range are more complicated than in the coaxial case because the guide wavelength is no longer that in free space, but depends on dimensions and dielectric constant in a rather complicated way. Also, the characteristic impedance depends on these, as well as on which of the various definitions one uses.

As in coaxial line, quarter-wave dielectric slugs may be used. For high power work, metal slugs of one form or another are preferable. One such is illustrated in figure 143.

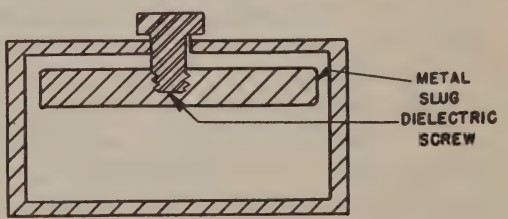


FIGURE 143.—Metal slug in wave guide.

Tuning screws.—A screw (or other rod) inserted from the broad face of the guide gives the effect of a shunt capacitance at that point. One form of screw which eliminates contact trouble by means of chokes is shown in figure 144. A fundamental difference between

stubs and tuning screws is that the former may introduce susceptances of either sign, whereas the latter is restricted to capacitive susceptances. This unfortunate limitation makes a $\frac{3}{8}\ \lambda$ spacing unusable. A screw inserted in the guide is a fairly sharp projection so would be certain to cause trouble at high power. This may be minimized by making the screw with a large diameter and a rounded end.

The simplest and most straightforward screw tuner is a single screw which may be moved along the guide a distance of at least half a guide wavelength. This enables one to match any given impedance to a line. The matching procedure is as follows: the screw is inserted a few turns and then the position of the screw along the line is adjusted for best match. Then the screw insertion is adjusted for best match, and the optimum position redetermined. These successive approximations rapidly produce a perfect match.

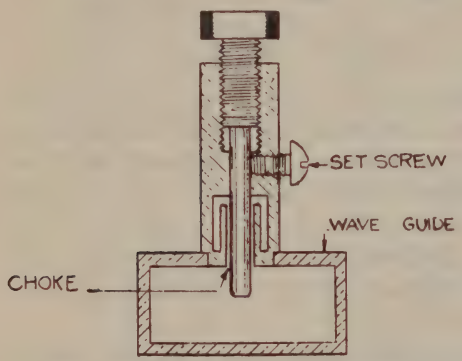


FIGURE 144.—3-cm. tuning screw, small size.

Tuners of this type are used considerably in low-power test-bench work where extremely well-matched loads are desired. They are very easy to tune up and quite insensitive to frequency changes. The tuning screw is mounted on a sleeve which fits around the wave guide and is moved along by a drive screw. Or, the drive screw may be omitted for quick rough adjustments. The tuning screw projects through a slot in the center of the broad face of the guide.

An arrangement which has been used considerably consists of three screws spaced at quarter guide-wavelength intervals. With this device, one can (theoretically) match any impedance to a wave guide. Practically, however, it is very difficult to match even a fairly small standing wave to a line. There are an infinite number of solutions to this three-screw problem, but finding even one of them is a difficult job. No series of successive approximations is certain to give the desired result. The best procedure is to use only two of the three screws, the center one and the proper one of the end ones. Rules for determining the proper screws have been described, but are complicated and not definitely proved to be valid. Once the proper pair of screws is determined, the solution is unique, but not a simple series of successive approximations. One screw must be adjusted off-optimum a bit and brought to match by the second as will be shown later. Considerable experience is needed to achieve satisfactory results.

For instance, consider the tuning of point A in figure 145. The load screw adds susceptance along the arc ABC'. As will be seen, point C' is the proper point for final matching. C' transforms to C at the input screw, $\lambda/4$ toward the generator. Then the input screw adds susceptance along the arc CY_0, give the desired matched line. Working backwards, we may determine the region which is tunable by these two screws. The input screw can match admittances along the semicircle CY_0. This semicircle is the transformation of the dotted semicircle $OC'Y_0$ at the load screw. Any point (such as A) lying in the shaded region may be transformed by adding susceptance with the load screw (along an arc such as ABC') to lie on the semicircle $OC'Y_0$.

Points in the unshaded region are not tunable with these two screws. However, going along the line a quarter wavelength either way transforms each of these two regions into the other one, so that the other pair of screws can tune the unshaded region of the diagram given.

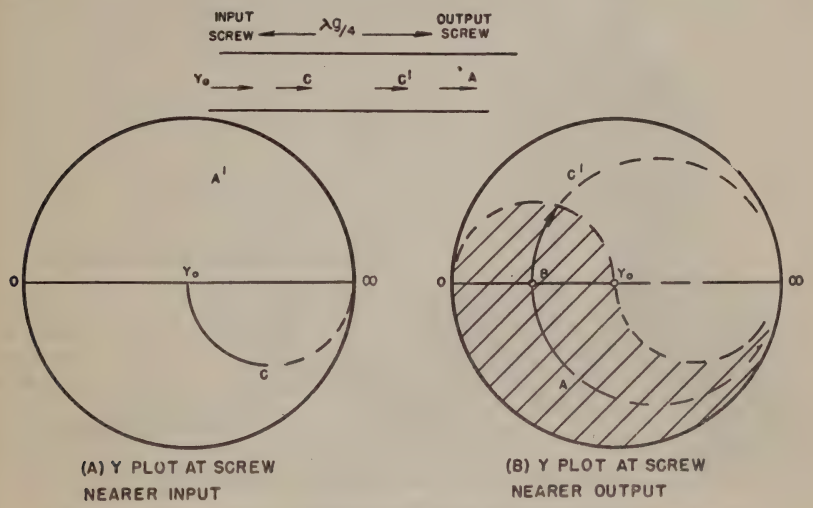


FIGURE 145.—Action of two $\lambda_g/4$ spaced screw tuners shown on an admittance diagram.

Consider again the example given above. Had we adjusted the load screw for best match we would have stopped at point B, the closest approach of arc ABC' to Y_0. The input screw would not then be able to improve on the matching of point B, since it would move one along the arc of a circle which has OB as a diameter. Had the input screw been tried first, it would have been found to make matters worse from the start, moving A' upward and to the right. Hence, the best one can do by successive approximations is arrive at point B. A proper over-insertion of the load screw (BC') is required before the input screw can achieve a match.

Figure 146 illustrates the appearance of some typical wave-guide tuners.

c. R-F Power Divider.

By the proper combination of branching lines and tuning stubs, one may construct a well-matched power divider, as shown in figure 147.

If the two outputs are matched and the stub plungers are ganged as shown ($l_1+\lambda/4=l_2$), the input will be matched at all settings of the plungers. When $l_1=\lambda/4$, stub number one will act as an ordinary stub support, allowing power to flow unimpeded into output number one. At the same time l_2 will be $\lambda/2$, which will present a short circuit across output number two and reflect all the power entering the number two branch. Since this short circuit occurs $\lambda/4$ from the input junction the number two branch will act as an ordinary stub support, allowing power to pass freely into branch number one.

When both stubs are lengthened by $\lambda/4$, the situation is reversed, and power flows into the number two output. At intermediate lengths

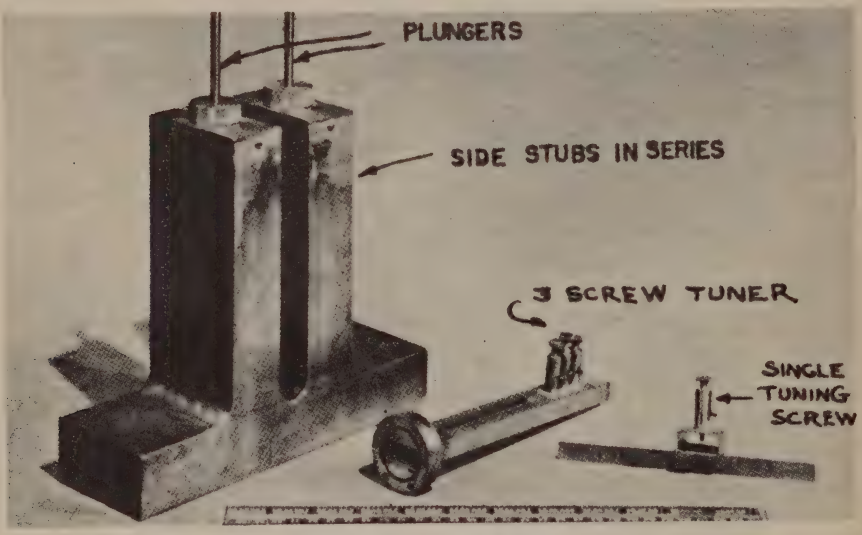


FIGURE 146.—Waveguide tuners.

of the ganged stubs l_1 will lie between $\lambda/4$ and $\lambda/2$, and l_2 will lie between $\lambda/2$ and $3\lambda/4$. This means stub one will have a capacitive susceptance varying from 0 to $+\infty$, while stub two will have inductive susceptance varying from $-\infty$ to 0. As may be verified mathematically, the combined admittance at the input T junction remains constant and equal to Y_0. The imaginary components due to the two stubs cancel, while the real parts of the two branches add to give Y_0. The power divides between the two outputs in the same ratio as the conductances of the branches. Calculation shows output one will vary as $\cos^2\theta$ (where θ is the angular increase in the electrical length of the ganged stubs) while output two varies as $\sin^2\theta$. This checks with our previous statement regarding match, since $P_1+P_2=\sin^2\theta+\cos^2\theta=1$.

The maximum attenuation of an output is limited by the ohmic losses in the λ/2 shorting stub. A calculation shows that this should limit one to about 60 db below full power in the case of ⅞-inch lines made of brass. An actual model showed about 50 db down to be the limit. This same model remained matched to r<1.5. A power divider is shown in figure 147.

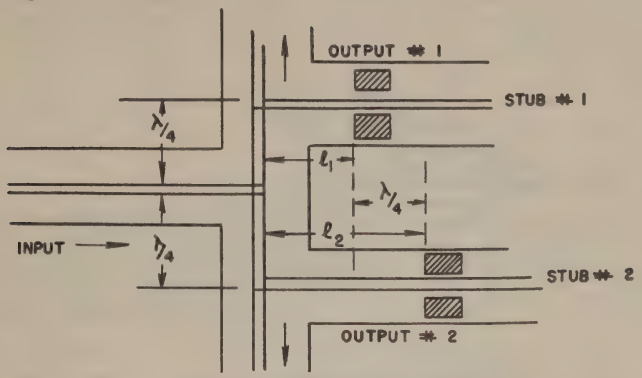


FIGURE 147.—Coaxial power divider.

4. WAVEMETERS AND ECHO BOXES

a. Wavemeters.

Types.—Wavemeters in common use are of three different types, namely (1) coaxial-line, (2) cylindrical-cavity, and (3) transition. Coaxial line and transition types are shown in figure 148, while cylindrical types appear in figure 149.

(*1*) *Coaxial line.*—A wavemeter of this type consists of a section of coaxial line small enough so that only the coaxial mode (*TEM*—also called $TM_{o,o,n}$) can exist. Resonance can then be obtained at odd quarter-wavelength points (1/4 λ, 3/4 λ, etc.) if the line is short circuited at one end and open at the other. Resonance can be obtained at even quarter-wavelength points (2/4 λ, 4/4 λ, etc.) if the line is short circuited at both ends. Loop coupling to the line is commonly used with the loop located near a short-circuited end so that it will always be in a position of maximum magnetic field. If more than one resonance point can be obtained on the wavemeter, it is self-calibrating since the distance between successive resonances is a half wave-length. If only one resonance point can be obtained, the wavemeter must be calibrated and it is important that the coupling loop be fixed in position if accurate results are to be obtained.

Type 401—10 cm. band.—Input and output coupling for either transmission or absorption use. Difference between two readings should be used for accurate work. Size: 1½ inches o-d, 21 inches long, 3-inch dial. Reading to 0.001 cm. with vernier. Made by Mico Instrument Company. See figure 150.

Type 402B—10 and 3 cm. bands.—One input coupling for absorp-

tion use. Difference between two readings should be used for accurate work. Size: ¾-inch o-d., 12 inches long, 2-inch dial. Reading to 0.001 cm. with vernier. Mico Instrument Company. See figure 150.

Hollingsworth or Type 1577—10 cm. band.—One input coupling for absorption use. Scale covers only one resonance point and is

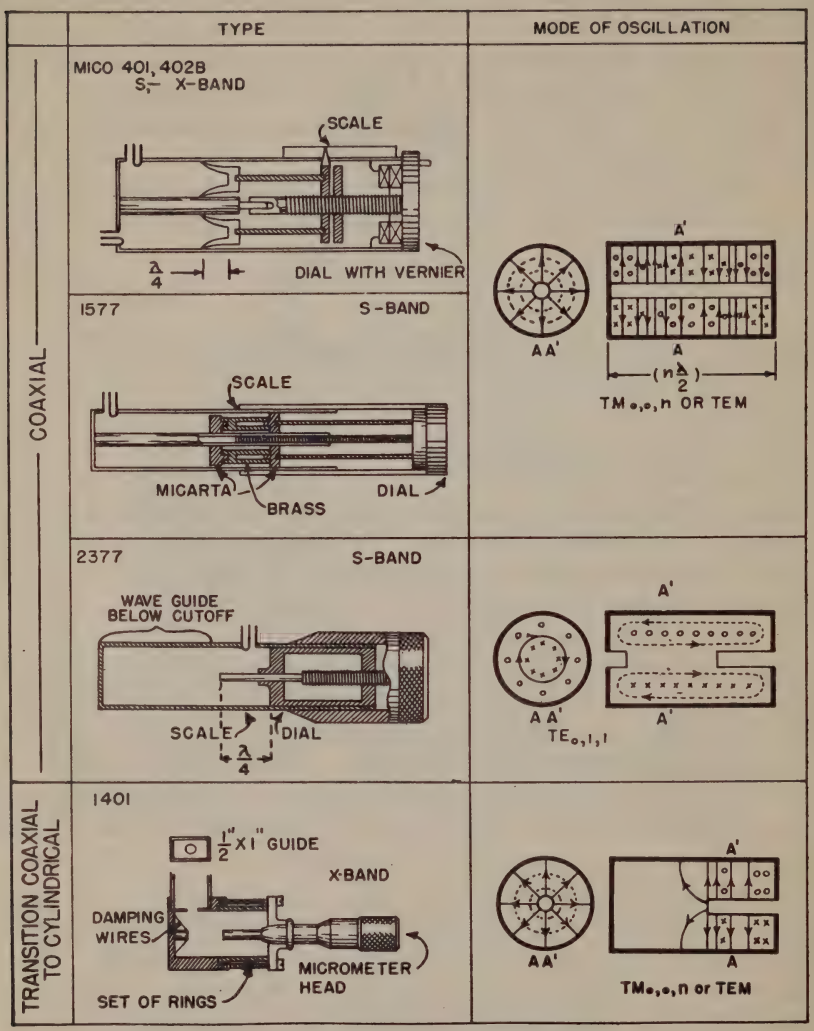


FIGURE 148.—Coaxial and transition types of wavemeters.

direct reading if zero set. Is accurate to within ±0.020 cm. within 1 cm. on either side of zero setting. Dial reads to 0.001 cm. See figure 151 (right).

Seely or Type 2377—10 cm. band.—Operates over linear portion of its characteristic and is direct reading if zero set. Is accurate to

0.1 percent within 1 cm. on either side of zero setting. Dial marked to 0.005 cm. Used as reaction instrument. Size: ¾-inch o-d., 6½ inches maximum length. See figure 151 (left).

(*2*) *Cylindrical cavity.*—This type of wavemeter consists of a resonant cavity. Its advantage over the coaxial type is that it has a higher Q and therefore better resolution. Its disadvantage is that

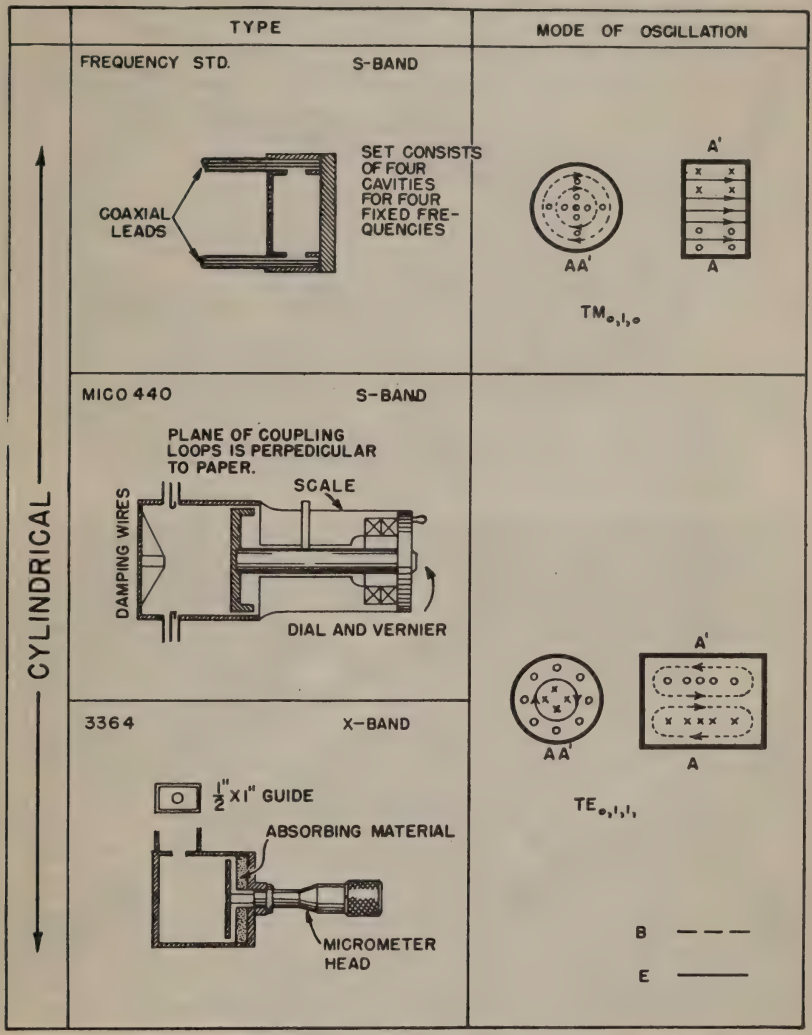


FIGURE 149.—Cylindrical types of wavemeters.

it is not absolute and, therefore, must be calibrated. It would be desirable to use the lowest possible mode so as to avoid confusion resulting from other modes. This is done with the fixed-frequency standards, but in order to have a variable cavity with a plunger at one end it is necessary to use a higher mode which has zero surface

currents at the point where the plunger meets the sides of the cavity. When a higher mode is used the lower modes are discouraged in various ways, e. g., by the method of coupling, by absorbing material behind the plunger, by absorbing material in grooves parallel to the surface currents on the end plate, etc.

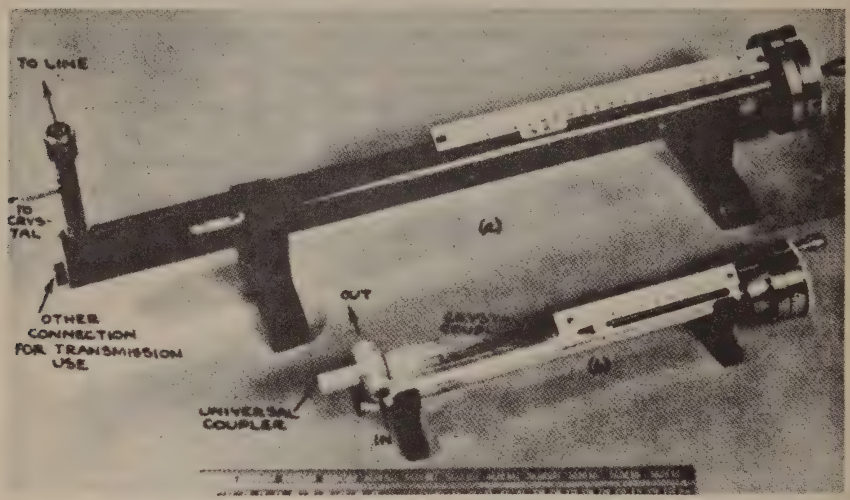


FIGURE 150.—Coaxial wavemeters. Left, type 401 wavemeter. Right, type 402-B wavemeter.

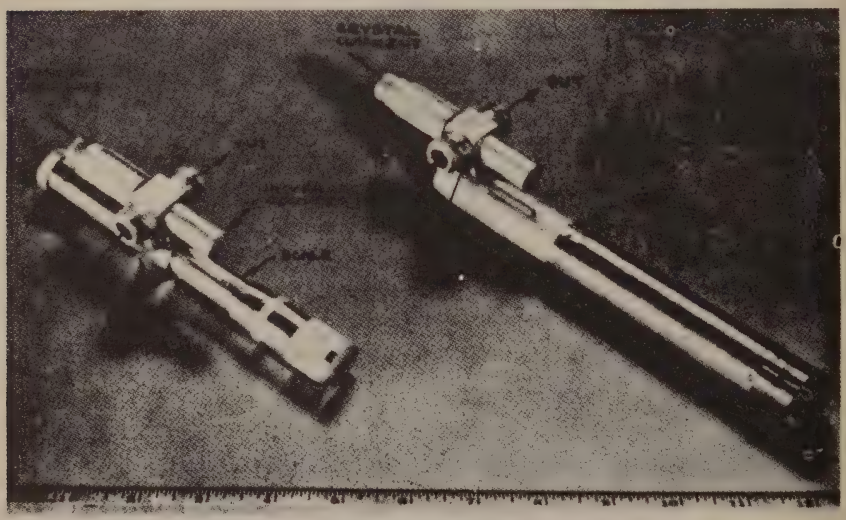


FIGURE 151.—Direct reading wavemeters. Left, Seely or type 2377 wavemeter. Right, Hollingsworth or type 1577 wave meter.

Type 440—10 cm. band.—Has a theoretical Q of 55,000, which gives a limit of accuracy of ± 28 kc/sec at 3,000 mc/sec. Experimentally a Q of greater than 40,000 has been obtained. Damping wires take out the undesired modes. Input and output couplings

require special coupling loops. Size: 6¼ inches o-d., 18 inches long, 4¼-inch dial with vernier. Mico Instrument Company.

Type 3364. Wide range 3 cm. band (3.16 to 3.43). If zero set it reads directly with no error at 3.20 cm. and at 3.40 cm. and with a maximum error of 0.009 cm. at 3.30 cm. The micrometer reading gives the wavelength starting with the second significant figure; i. e., if m is the micrometer reading $\lambda = \left(3\frac{+m}{1000}\right)$cm. Absorbing material in the back cavity reduces the Q of undesired modes. One must be very careful not to be misled by these low Q modes, ½-inch x 1-inch

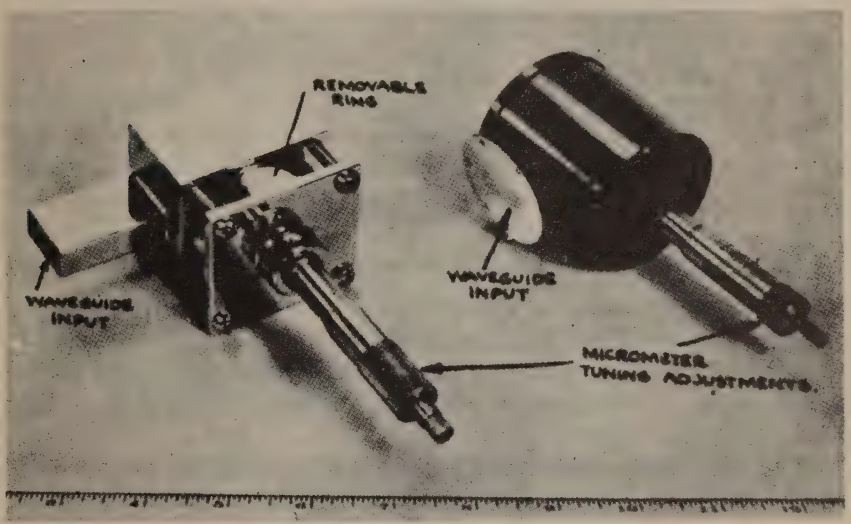


FIGURE 152 —3 cm. wavemeters. Left, Madam X or type 1401 wavemeter. Right, wide range type 3364 wavemeter.

wave guide input with iris coupling is provided. $Q_0 \sim 25{,}000$. Size: 2¼-inches o. d., 2⅜-inches long plus micrometer head. See figure 152 (right). Auto-Ordnance Corp.

Frequency standards—10 cm. band.—Sets of four fixed cavities are calibrated in Group 55 of the Radiation Laboratory and sent to industrial laboratories for checking wave meters. They have coaxial inputs and are used as transmission instruments. Each set is supplied with lossy cable, crystal, and type N-connectors. The calibration system is as shown in figure 153.

(*3*) *Transition.*—This type of wave meter can be thought of as half coaxial line and half cylindrical cavity. It consists of a cylindrical cavity along whose axis projects a piston whose diameter is considerably smaller than that of the cavity. The diameter of the cavity is large enough so that the $TE_{0,1,1}$ mode may be excited in the cavity. This is not the simple $TE_{0,1,1}$ mode when the plunger is inserted and

all dimensions of both the cavity and the plunger determine the wavelength. Consequently, this wavemeter must be calibrated.

Madam X or Type 1401–3 cm band.—Size of cavity variable by means of six interchangeable rings which cover overlapping regions of the band. Each ring covers about 0.12 cm. Scale is not direct reading and a calibration curve is necessary. Damping wires are used

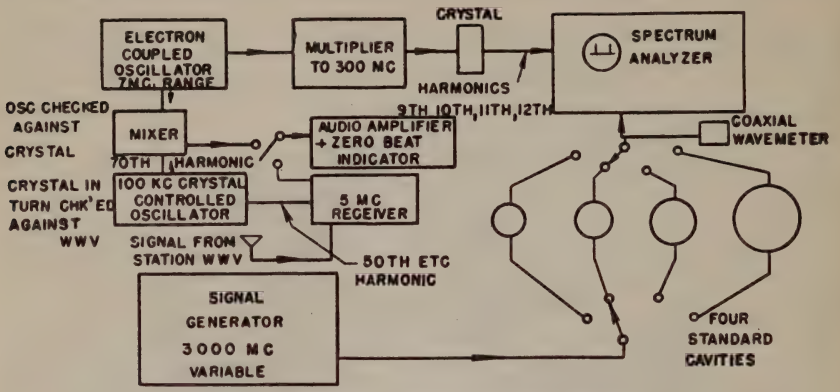


FIGURE 153.—Calibration system for frequency standards.

to damp lower modes. ½-inch x 1-inch wave guide inputs with iris coupling. See figures 152 (right) and 148.

Methods of use.—Depending upon whether it is in the main line or on a side arm, a wave meter is said to be used as a "transmission" or a "reaction" instrument.

(*1*) *Transmission instrument.*—A typical example of the use of a wave meter as a transmission instrument is illustrated in figure 154.

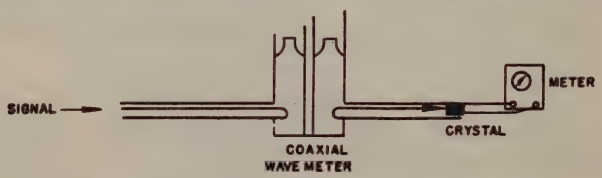


FIGURE 154.—Wave meter as a transmission instrument.

Maximum power is coupled through the wave meter at resonance. Therefore, wave meter readings are taken when the meter deflection is a maximum.

(*2*) *Reaction (or absorption) instrument.*—A typical example of the use of a wave meter as reaction (or absorption) instrument is shown in figure 155.

The distance L_1 is such that the wave meter arm presents a high impedance at the junction with the main line when the wave meter is not resonant. At resonance, however, it presents a low impedance so that less power gets to the load and to the crystal. The crystal branch could be placed elsewhere if it were so desired. For example, it might

be placed on the wavemeter branch if it is placed at a point of normally (i. e. off-resonance) high impedance on that branch. Also it might be placed anywhere on the main line between the wavemeter branch and the load. These alternative absorption set-ups are shown in figures

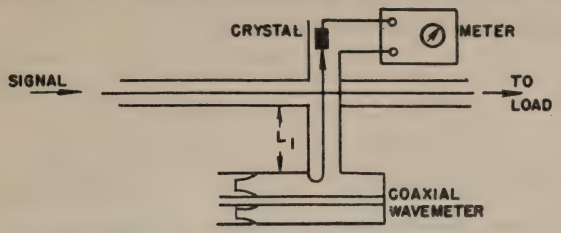


FIGURE 155.—Wave meter as a reaction instrument.

156 (A) and (B). In (A) the length L_1 must be such as to present a high impedance looking toward the wave meter at the junction with the crystal branch. Length L_2 is independent of L_1, for when the wave meter is not resonant, the impedance at the junction with the

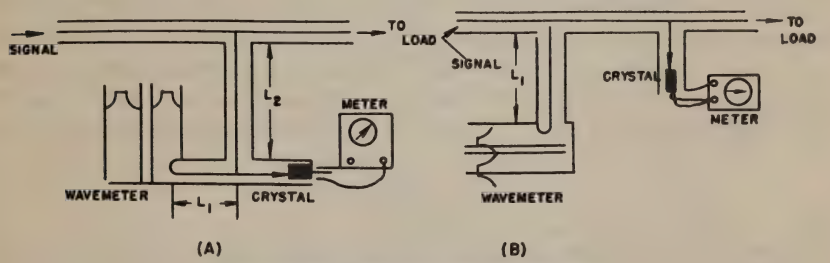


FIGURE 156.—Alternative absorption set-ups.

main line depends only on the impedance of the crystal and the length of line up to it.

Coupling sections of the cross and T types with type N fittings have been made which fit on several different wave meters. (See figure 157.)

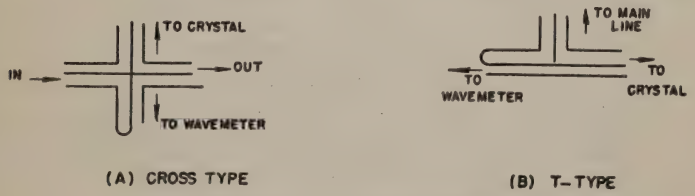


FIGURE 157.—Wave meter couplers.

For example, the universal coupler fits on the Seely, Hollingsworth and 402B wavemeters, as shown in figures 150 and 151. The type 2274 coupler, which is only slightly different from the universal, fits on both the 401 and 402B wave meters.

In the absorption case the current from the monitoring crystal should vary as shown in figure 158 (A) as the wave meter setting is changed. However, when the coupling loop is too large, the crystal current may vary as shown in figure 158 (B). This coupling is too tight and can be loosened by rotating the loop. The rotation serves to decrease the effective area of the loop.

FIGURE 158.—Current through monitoring crystal as wavemeter is tuned.

The accuracy of a wavemeter depends upon the accuracy with which its scale can be read and the loaded Q. For example, if accuracy of ± 0.001 cm. is desired, the scale must be fine enough to read at least 3 decimal places and the resolution of the wave meter must be great enough to determine when the tuning is 0.001 cm. or less from resonance. In other words, the frequency response curve or the wave meter must drop to its "half power" points within 0.001 cm. of the resonant wavelength. Therefore at 10. cm. the loaded Q must be

$$Q_L = \frac{\lambda}{\Delta\lambda} = \frac{10.0}{2(0.001)} = 5{,}000$$

The coupling loops used with a wavemeter must be small enough to avoid serious reduction in loaded Q.

b. Echo Boxes.

Use.—An echo box is a high Q resonant cavity which receives r-f energy through a pick-up antenna during the transmitted pulse and reradiates this energy through the same antenna immediately after the pulse. The reradiated energy is picked up by the radar set. Since this energy from the echo box dies off exponentially, it will appear on an A-scope indicator as an exponential curve. The time from the end of the transmitted pulse to the time that the echo box signal is lost in noise is called the "ringing time" of the echo box. A crystal may also be coupled into the echo box to indicate the tuning of the box and the amount of energy reaching the box. The crystal current is an indication which can be counted on in case the receiver is not functioning properly. Also the crystal may be calibrated to measure transmitter output for a standardized position of the pick-up antenna. A typical set-up is illustrated in figure 159.

At first the exponential signal is so strong that the receiver is satu-

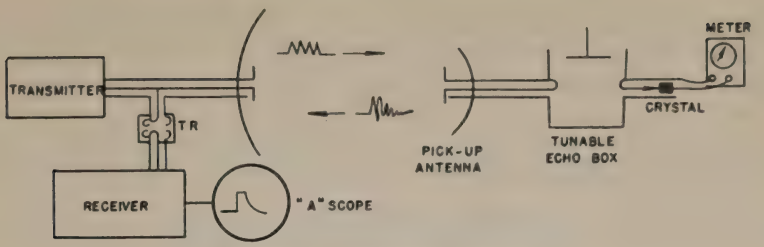


FIGURE 159.—Typical echo box set-up.

rated. As a result the pattern on the A-scope looks as indicated in figure 160.

Types—*(1) Tuned Echo Boxes.*—A tuned (or tunable) echo box is of the sort indicated in figure 159, i. e., it consists of an adjustable cavity operating in a single mode. If calibrated, the setting of the plunger at resonance will indicate the wavelength. Tuned echo boxes operating in the $TE_{0,1,1}$ mode are made to cover the 10 cm. band in two steps. (See figure 161.) The S/A box tunes over the S_A and S_S bands, and the S/G box tunes over the S_S and S_G bands.

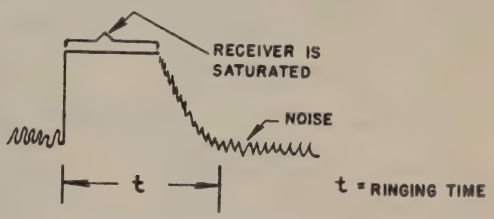


FIGURE 160.—Echo box signal on A-scope.

The overall performance of a system can be checked with an echo box as follows: Measure the ringing time on two different occasions.

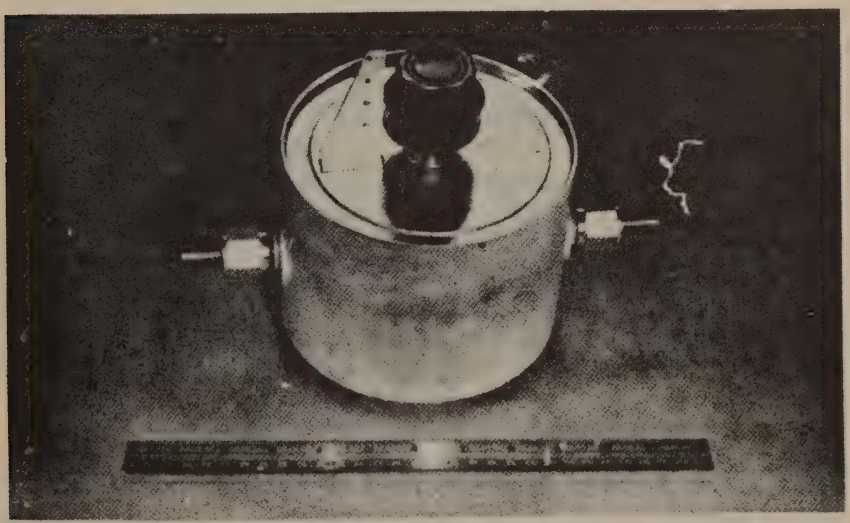


FIGURE 161.—10-cm. tunable echo box.

Suppose that the second ringing time is greater than the first by Δt seconds. Then

$$10 \log_{10}\left(\frac{P_2}{P_1}\right)=\frac{10}{A}\Delta t$$

where $10 \log_{10}\left(\frac{P_2}{P_1}\right)$ is the increase in transmitted power, decrease in noise power, or change in antenna gain, etc., or a combination of these.

$$A=2.3\frac{Q}{2\pi f}$$

Q=loaded Q of echo box

f=operating frequency in cycles per second.

Calibrated echo boxes may be used to analyze the spectrum of the transmitter. The Q of an echo box is great enough to permit selection of the components of a transmitter spectrum. Therefore the transmitter's spectrum can be obtained by plotting wavelength versus crystal current as the echo box is tuned.

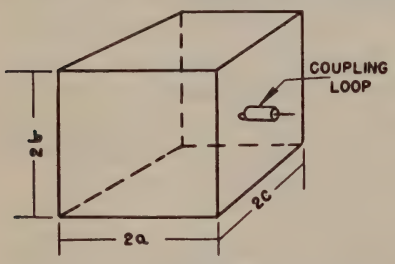


FIGURE 162.—Rectangular box type fixed-tuned echo box.

Experimental echo boxes for the 3-cm. band designed to operate in the $TE_{0,3,5}$ mode are being tested.

(*2*) *Untuned echo boxes.*—An untuned echo box is a fixed cavity large enough so that there is at least one mode within the bandwidth of the receiver and within the spectrum of the transmitter operating at any wavelength within a certain band.

The $TE_{l,m,n}$ and the $TM_{l,m,n}$ modes which can exist in a rectangular box (figure 162) are determined from

$$\frac{4}{\lambda}=\sqrt{\left(\frac{l}{a}\right)^2+\left(\frac{m}{b}\right)^2+\left(\frac{n}{c}\right)^2}$$

where l, m, and n are integers. For any one set of l, m, and n the following modes are possible:[1]

TE	TM
l, m, n	l, m, n
l, n, m	l, n, m
m, l, n	m, l, n
m, n, l	m, n, l
n, l, m	n, l, m
n, m, l	n, m, l

There are as many of these groups of modes as there are different sets of integers which fit the above equation. The equation also

[1] This is not strictly true. Some zero values are forbidden.

shows that if the cavity is a perfect cube ($a=b=c$), there are a large number of combinations of the quantum numbers (l, m, and n) which lead to the same value of λ. This fact indicates that in a perfect cube there may be 12, 24, or more states or modes for just one wavelength. This duplication is known as degeneracy. Shifting from a cube to a rectangular box ($a \neq b=c$; i. e., square on one end) spreads the modes, leaving only a two-fold degeneracy. Irregularities in sides such as bulges and dents probably tend to break up this remaining degeneracy.

The main objection to these echo boxes at 10 cm, is their awkward size. Since the Q for a particular mode is proportional to a/λ), where a is a typical side, a satisfactory untuned echo box at 10 cm, has a linear dimension on the order of a meter. An equivalent box at 3 cm. would be about one third as big.

Boxes in use, 10-cm. band.—The Radiation Laboratory has a 3-foot near-cube copper resonator. The theoretical Q is approximately 450,000. Ringing times of over 100 microseconds are obtained with 35 kw. peak power on a 1 microsecond pulse system. The box has one coupling loop with type N fittings.

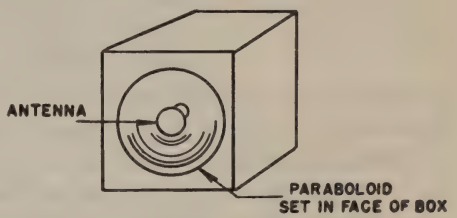


FIGURE 163.—3-cm. untuned echo box.

Boxes in use, 3-cm. band.—Sperry Gyroscope Co., has designed a 13-inch near cube box. (See figure 163.)

This box is placed in front of a radar set for overall checks. For some points in the frequency band it is necessary to push in the sides to get a large enough ringing time. A typical ringing time for this box is 15 microseconds.

5. TEST BENCH R-F OSCILLATORS

a. General Purpose Generators

Any tube described in the section on local oscillators is also suitable for use as a signal generator for laboratory measurements. The 417A, 707A, and 723A/B are commonly used for this purpose. When the generator is to be used with an audio amplifier as an indicator of rectified r-f current, audio-frequency amplitude modulation is most conveniently accomplished by applying a square voltage wave to the reflector electrode.

High output generators.—

417, 417A. The 417A reflex Klystron can be operated with power inputs up to 75 watts. This is most efficiently done at 1,200 volts and 60 ma., obtained with about +50 volts grid bias. Maximum power output is about 1 watt. A blower must be used to cool the tube. Modulation can be accomplished by applying a square wave to the

reflector, or by applying a sine wave of about 50 volts r. m. s. to the control grid (using zero grid bias). To obtain the maximum power output a double-stub tuner or similar device should be inserted in the r-f line as near the tube as possible, and with no lossy cable in between.

412 and 710. The 412 and 710 are two-cavity Klystrons made by the Sperry Co. in very small quantities. The tubes cover wavelength ranges of approximately 8.5 to 9.6 and 9.5 to 11.5 cm., respectively. The Sperry model 11C tuner tunes both cavities with a slight adjustment over a wavelength range of about 3 mm. Initial alignment of the tubes requires special apparatus and skill. Oscillation occurs at a number of voltages between 300 and 2,000 volts. Maximum cathode to anode voltage 2,000 volts. Maximum power input 100 watts. Typical low-voltage operation:

anode volts=450 v.
grid volts=+30 v.
anode current=20 ma.
grid current=2 ma.

A blower must be used to cool the tube for inputs of over 20 watts. Power outputs of 1 or 2 watts may be obtained.

410R. The 410R is similar to the 412 and 710 except that a set of fins is attached to the anode to dissipate heat. Maximum power input 300 watts. Maximum voltage 3,000 volts. Pin 5 on the base is a focusing electrode which is run at cathode potential or up to 100 volts negative. Typical operating conditions are listed in the following table:

Volts	Ma.
900	35
1500	75
2400	150

Maximum power output is 20 watts. This tube is made by Sperry in small quantities on special order.

419. The 419 is a 3 cm. reflex Klystron, made by Sperry in small quantities, which produces 100 to 200-milliwatts of r-f power. The tube is mechanically similar to the 417. The r-f output lead terminates in a coaxial connector which must be connected to a flexible cable. A Sperry model 12 tuner is used to support the tube, and tune over a wavelength range of about 2 mm. with a single fine adjustment. By adjusting the three coarse screws the range from 2.9–4.5 cm. can be covered.

The standard 419 contains radial type resonator grids, and can be operated at 1,100 volts maximum. The corresponding plate current is 30–40 ma. Pin 5 is a focusing electrode run at cathode potential.

The characteristics of a typical tube at 1,050 volts are tabulated below.

Reflector voltage	Power output (milliwatts)	Frequency shift in mc. between half-power points	Frequency shift with reflector voltage (mc/v)
−80	15	40	2.2
−160	50	24	0.8
−300	100	18	0.4

Generators for wide wavelength ranges.—

707A. This versatile tube is commonly used with a wave-guide-type cavity tunable from 8–12.5 cm. by adjustment of two plungers. The above cavity is made for Radiation laboratory by Humble Oil Co.

A coaxial cavity employing higher harmonics enables the 707A to be used continuously from 7–30 cm. Measurements with such a cavity $\frac{5}{4}\ \lambda$ long show that the electrical tuning is reduced to 3 mc. between half-power points, in the 10-cm. region.

1365-XO. A 723A/B equipped with an extra tuning strut for wider range. Approximate wavelength range 3.15–3.8 cm. One tube went to 4.4 cm. Made by Western Electric in experimental quantities. The tuning mechanism is not mechanically good.

1332-CT and 1280-CT. These two tubes are respectively three- and two-gap single resonator velocity-modulation oscillators employing magnetically focused electron beams, and developed by A. L. Samuel of the Bell Telephone Laboratories. The use of wave guide type cavities makes possible wide tuning ranges. The extreme limits between which it is convenient to tune a single cavity are approximately 5.5–15 cm. for the 1332–CT, and 3–8 cm. for the 1280–CT. Western Electric is planning to make 100 of the latter tube for experimental work. No wide range cavity has yet been designed for the 3–8 cm. range.

1289-CT. Three gap velocity-modulated oscillators which works in 3 to 8 centimeter range. Bell Laboratories have designed a wide range 3–8 cm. cavity for this tube.

b. Modulated Systems.

When it is desired to use an amplifier as the indicator on a low-level standing wave system it is necessary to provide modulation of the c. w. oscillator. In the past this has normally been done by providing sine-wave modulation. Since the oscillator is in most cases a reflex Klystron, the modulation is accomplished by modulating the cavity, the reflector electrode, or the electron beam focusing potential. The cavity is seldom modulated because of the additional power required to do so. Sinusoidal modulation of the reflector

or grid produces frequency modulation as well as amplitude modulation; in general the total amount of frequency modulation produced at 3 cm. is greater than that produced by the same method at 10 cm. Since this frequency modulation is undesirable in either case, a square-wave generator is now often used to provide the modulation. To understand how this modulation is produced examine figure 164 which shows the frequency and power output of a typical velocity modulation oscillator as functions of reflector voltage.

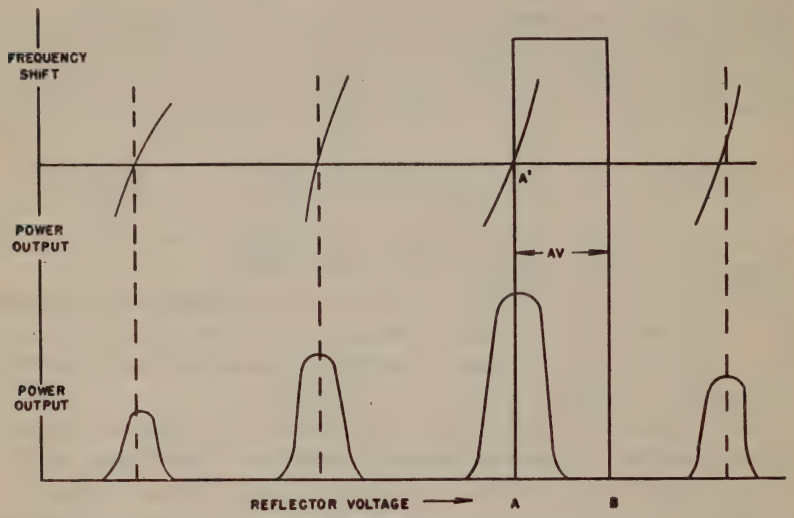


FIGURE 164.—Frequency shift and power output as functions of the reflector voltage for a typical velocity modulation tube.

If the tube was operating without modulation at point A and at a frequency represented by point A', and if a square wave of amplitude ΔV is imposed on the reflector, the potential of the reflector is equal to A during one-half of the cycle and equal to $A+\Delta V=B$ during the other half. At point B the tube is not oscillating, while at A the tube is oscillating, and the power output and the frequency is constant. In actuality since the wave front of the square wave is not infinitely steep, some slight frequency modulation does occur, but it is not enough to disturb the measuring system. The signal will of course have the normal amplitude-modulation spectrum.

c. Power Supplies.

The frequency of a velocity-modulated oscillator tube is a function of the cavity and reflector voltage applied to the tube. Therefore, in order to obtain the frequency stability needed for test-bench oscillators, it is necessary to use an electronically regulated voltage source.

The circuits of various supplies in use at the Radiation Laboratory are shown in the following figures:

Figure 165 shows a power supply suitable for 417, 707, or 723A/B oscillators. Figures 166 and 167 give adapter connections for the 417 and 707 tubes.

Figure 168 is a schematic of a power supply and modulator suitable for the 707A and 723A/B tubes. The modulator is a multivibrator square-wave generator (6SN7) and limiter (6V6–GT) and is connected to the reflector lead.

Figure 169 illustrates a power supply for a 417 Klystron, and figure 170 shows a power supply for a 419 Klystron.

6. POWER MEASUREMENTS

a. High-Level Power Measurements.

Several methods of measuring power at high level in the microwave region are now available, namely, (1) continuous flow calorimeter, (2) Johnson type power measurer, and (3) low-level power measurer with attenuation.

(*1*) *Water loads.*—Of the methods listed above, the continuous-flow calorimeter is probably the simplest to use and easiest to build. The construction of such a device for use as a coaxial transmission line is shown in figure 171.

A short section of coaxial transmission line is short-circuited at the far end and closed at the other end by means of a thick dielectric bead at *C*. The length of the device should be such that the distance from the bead to the short circuit is several wavelengths. This insures complete absorption of the radio-frequency energy. The bead at *C* is made to be ¼ λ long at the frequency to be measured and the dielectric constant of the bead is chosen such that—

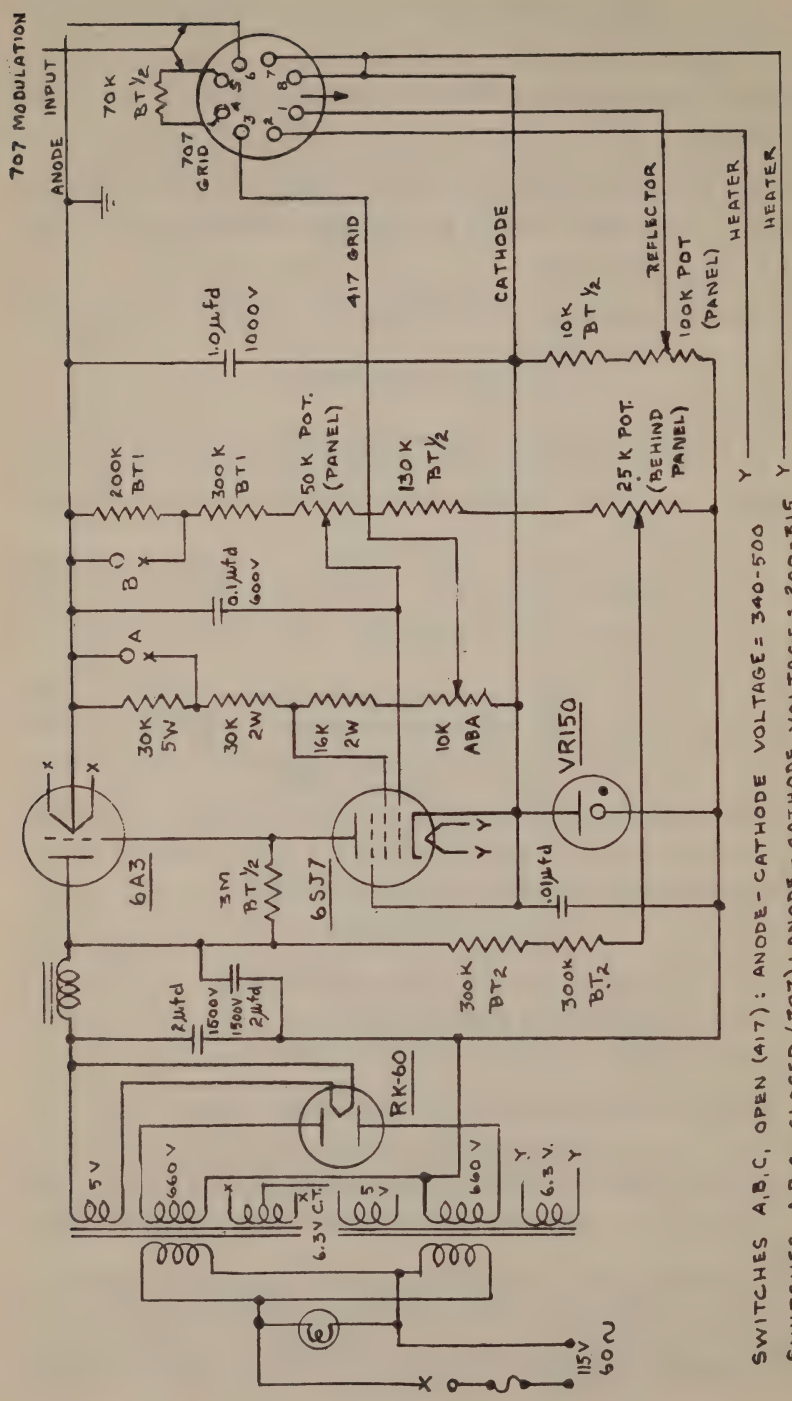
$$Z_d = \sqrt{Z_{air} \times Z_{H_2O}}$$

Z_d = impedance of the dielectric-filled line
Z_{air} = impedance of the air-filled line
Z_{H_2O} = impedance of the water-filled line

Glass and micalex have been used as the dielectric.

Since the bead is ¼ λ long at only one frequency, the device is somewhat frequency sensitive and it is necessary to provide matching transformers if the device is to be used at some other than the designed frequency.

This instrument is of the constant-flow calorimeter type; i. e., water flows continuously through the chamber and is heated by the r-f energy. Thermocouple leads are placed in the input and in the outlet to measure the rise in temperature of the water in flowing



SWITCHES A,B,C, OPEN (417): ANODE-CATHODE VOLTAGE = 340-500
SWITCHES A,B,C, CLOSED (707): ANODE-CATHODE VOLTAGE = 200-315

FIGURE 165.—Type M-2 power supply.

707

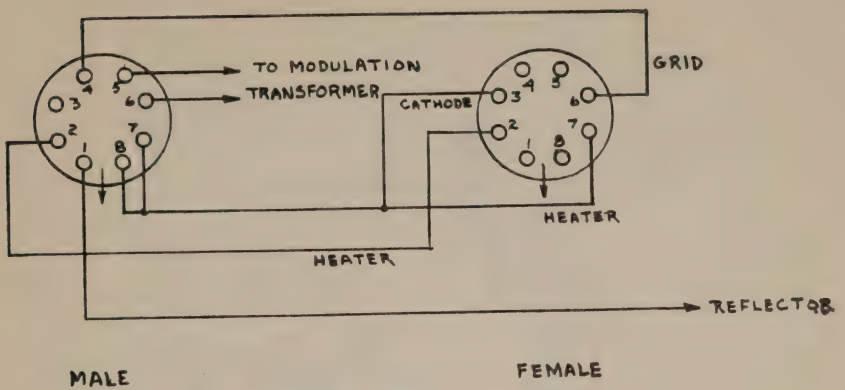


FIGURE 166.—Cables for M-2 supply (707).

417

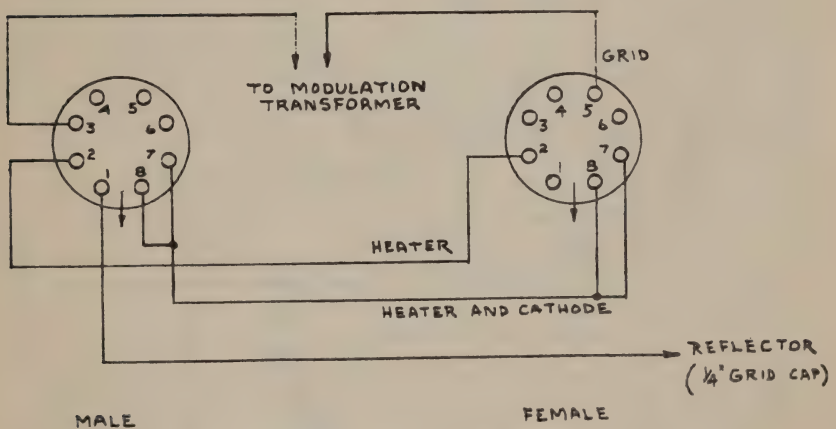


FIGURE 167.—Cables for M-2 supply (417).

through the cavity. To find the power then it is necessary to measure the rate of flow of the water and the change in temperature of the water.

If

m = mass of water flowing in unit time
w = power in watts absorbed in the load
k = heat capacity of water in calories per gram per degree = 1.0 cal/g/° C.
ΔT = change in temperature of the water
J = factor converting calories to joules, 4.19 joules = 1 cal.,

then

$$w = Jkm\Delta T.$$

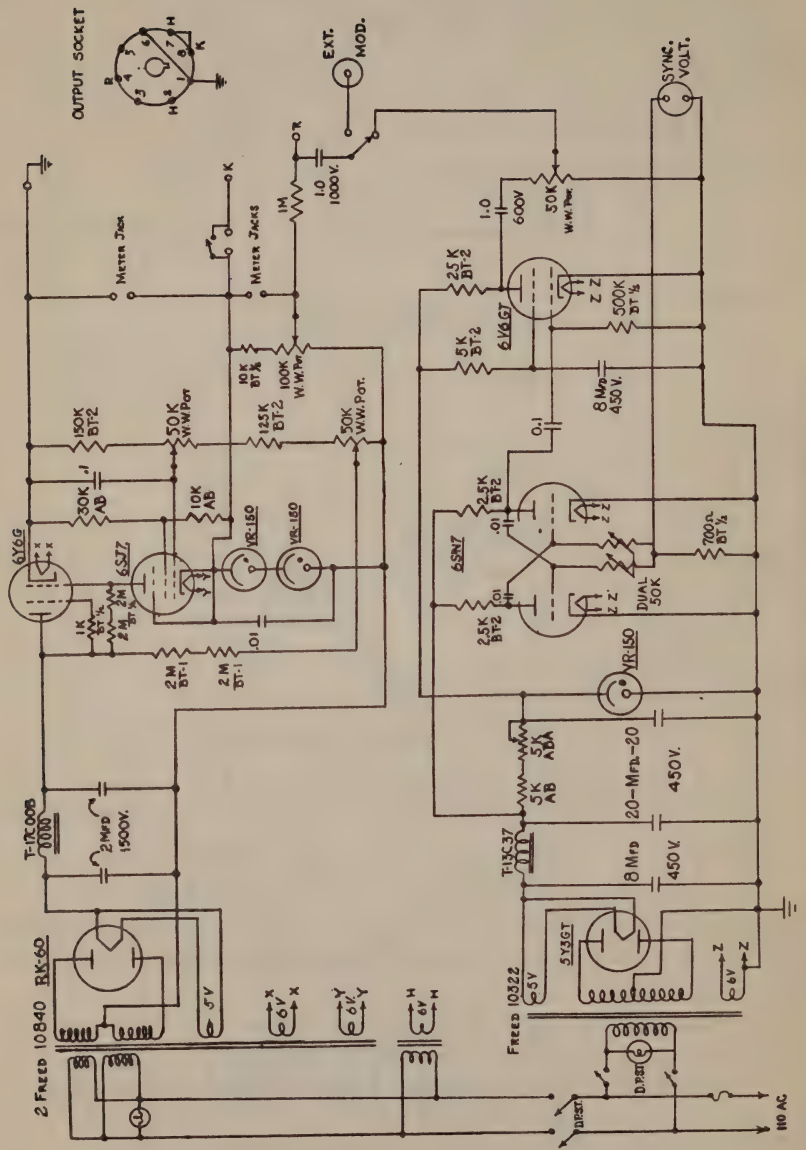


FIGURE 168.—Type N power supply and modulator.

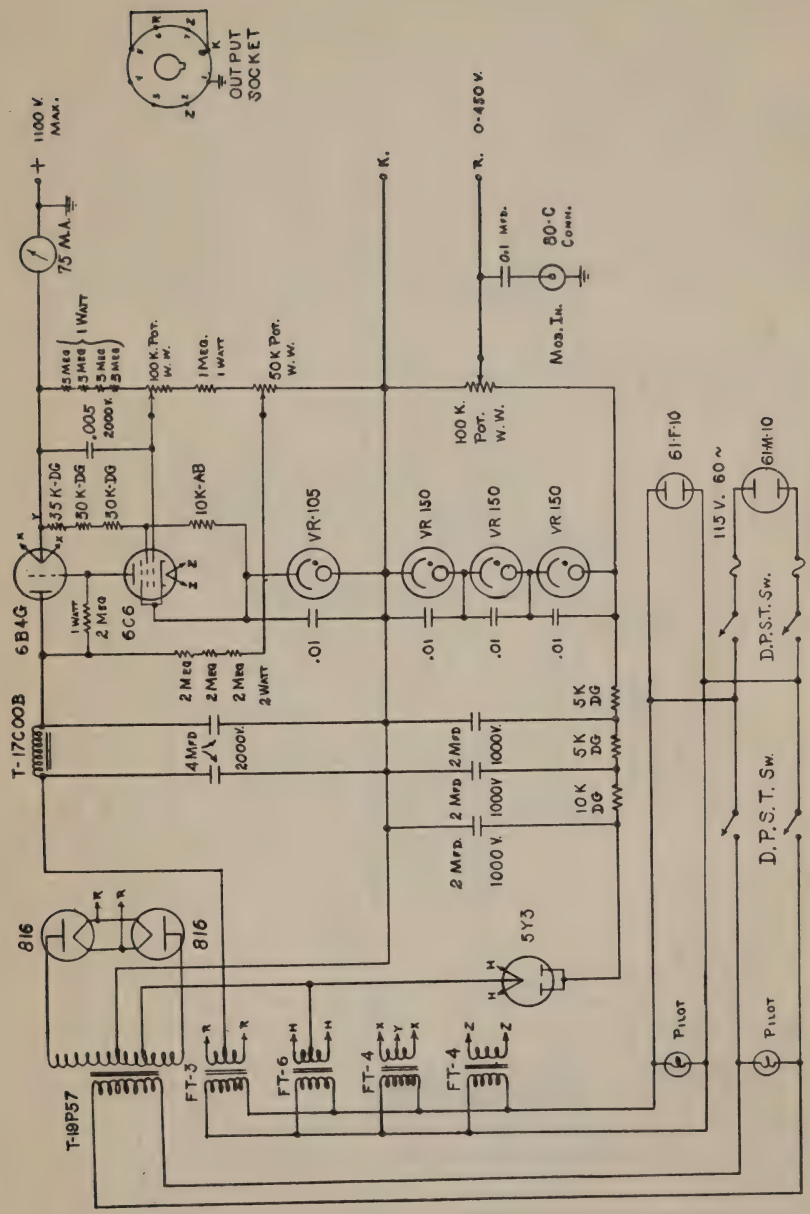


FIGURE 169.—Power supply for 417.

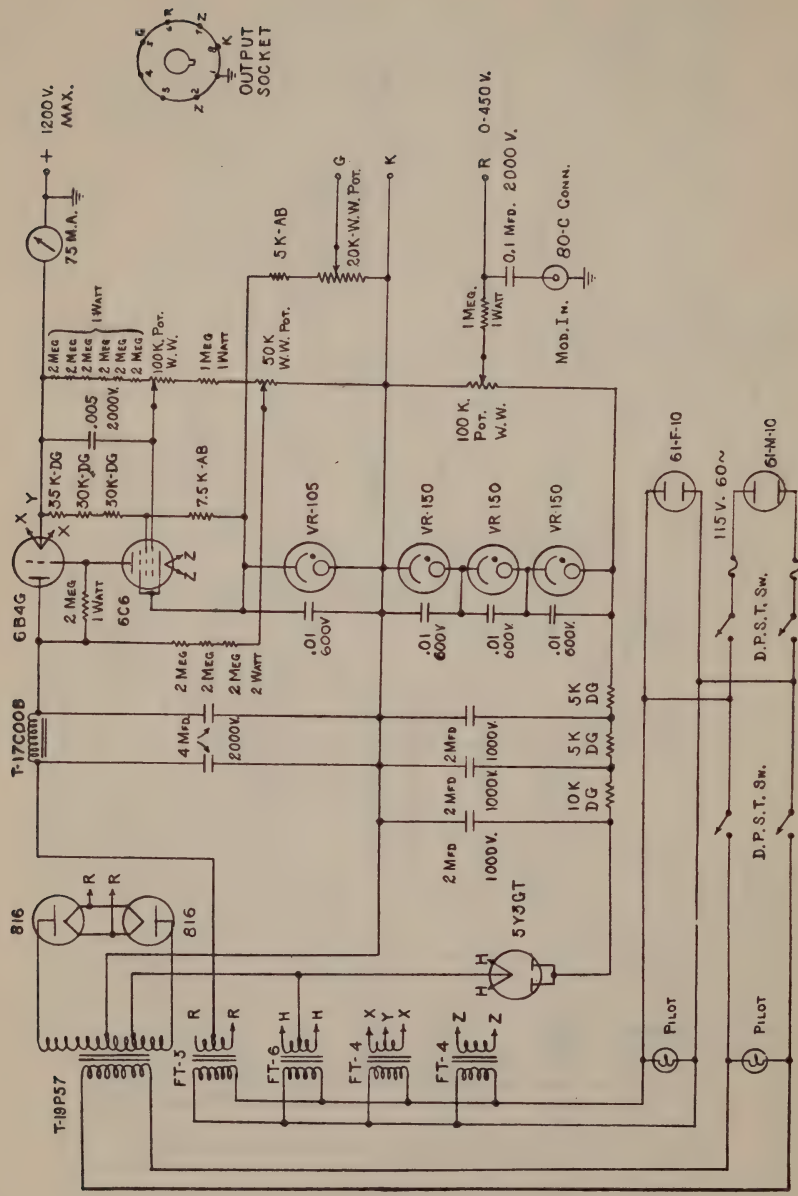


FIGURE 170.—Power supply for 419.

If the density of water is ρ then

$$m=\rho v$$

where v is the volume of water flowing through the cavity in unit time and

$$w=Jk\rho v\Delta T=4.19v\Delta T$$

ΔT is not measured directly but rather the current in microamperes which is in practice a linear function of the difference in temperature of the two thermocouple junctions. It is necessary, therefore, to know the calibration constant for the thermocouples and meter. This constant can be obtained simply by placing the thermocouple junctions in water baths which differ in temperature by ΔT_0 and noting the deflection d of the meter used.
Thus

$$w=Jk\rho v\frac{\Delta T_0}{d}D=4.19v\frac{\Delta T_0}{d}D$$

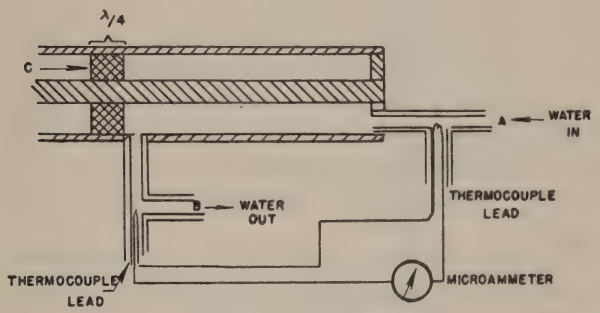


FIGURE 171.—Coaxial water load.

where D is the deflection of the meter during the power measurement. Since J, k, ρ, and $\frac{\Delta T_0}{d}$ are constants,

$$w=AvD$$

where

$$A=4.19\frac{\Delta T_0}{d}$$

Thus once the constant A is known, the power absorbed in the termination can be obtained by multiplying it by the rate of flow of the water, v, and the deflection of the meter, D.

This method of measuring has the advantage of being absolute. The rise in temperature can be made small, thus minimizing the energy lost by radiation and conduction down the line. The cooling correction is small enough that it can be neglected. In order to maintain a constant rate of flow, however, it is necessary to use some device to supply water at a constant head. The most serious disadvantage is the fact that such a device must terminate a line and therefore absorbs

100 percent of the power and it cannot be used to monitor power in a transmission line (except by mechanically switching it in and out of the circuit).

The water load for a wave guide is electrically the same as that for a coaxial line; that is, it consists of a water chamber which terminates the guide and absorbs all of the power. It is matched to the guide by a quarter-wave-length transformer. This transformer consists of a section of guide whose electrical length is one-quarter wave length and which is filled with a material whose dielectric constant is the geometric mean of those of air and water. (See figure 184.)

(*2*) *Johnson meters.*—Another type of power measurer, which overcomes the greatest draw-back of the water termination, is the Johnson-type wattmeter. This device requires only 1 percent of the energy in the transmission line instead of 100 percent as is the case with the water load. Figure 172 shows the construction of such a device for use in a wave-guide transmission line. A small section of the wave guide wall is removed and replaced with a section of a high-resistance material, namely, a thin (0.015-inch) sheet of constantan. It can be shown that at magnetron power levels enough heat is dissipated in the constantan to produce appreciable temperature changes. Since the conductivity of the wave guide is greater than that of the constantan there will be maintained a difference in the temperature of these two materials. This difference in temperature can be used to give an indication of the power.

In order to measure this difference in temperature a platinum resistance thermometer is used; 200 ohms of 0.002-inch platinum wire is wound on the constantan section of the wave guide and another 200 ohms is wound on the guide. These two windings serve as two of the arms of a Wheatstone bridge, as shown in figure 173.

If the bridge is balanced initially, any change in temperature of one of the coils with respect to the other will unbalance the bridge, due to the change in resistance of the platinum with temperature. The Johnson-type meter is compensated for changes in ambient temperature because the two windings are on a section of the transmission line and any change in ambient temperature will affect both coils to the same degree. The thermal capacity of the active element is made as small as possible in order to speed up the response. Readings cannot be taken instantaneously with either the water load or the Johnson meter, because there is a time delay between the turn-on of the power and thermal equilibrium of the measuring device. However, readings can usually be taken between 1 to 3 minutes after the turn-on.

The unbalanced current in the bridge is directly proportional to the power in the line. Thus a straight line calibration curve can be drawn by measuring only one point. The accuracy is to within 5 percent. The instrument can be calibrated against a water load or by passing a

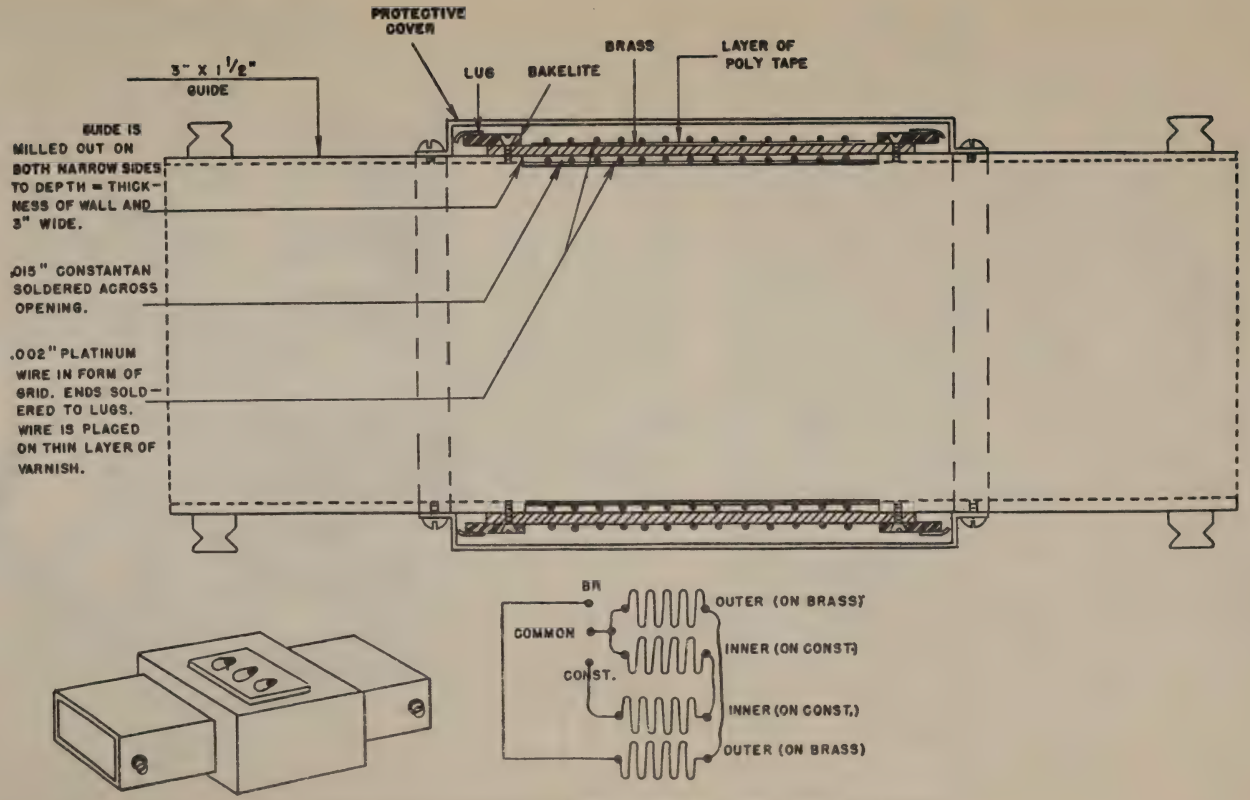

PROTECTIVE COVER
3" X 1 1/2" GUIDE
LUG
BAKELITE
BRASS
LAYER OF POLY TAPE
GUIDE IS MILLED OUT ON BOTH NARROW SIDES TO DEPTH = THICKNESS OF WALL AND 3" WIDE.
.015" CONSTANTAN SOLDERED ACROSS OPENING.
.002" PLATINUM WIRE IN FORM OF GRID. ENDS SOLDERED TO LUGS. WIRE IS PLACED ON THIN LAYER OF VARNISH.
BR
COMMON
CONST.
OUTER (ON BRASS)
INNER (ON CONST.)
INNER (ON CONST.)
OUTER (ON BRASS)

FIGURE 172.—Construction of Johnson meter in 1¼-inch x 3-inch wave guide.

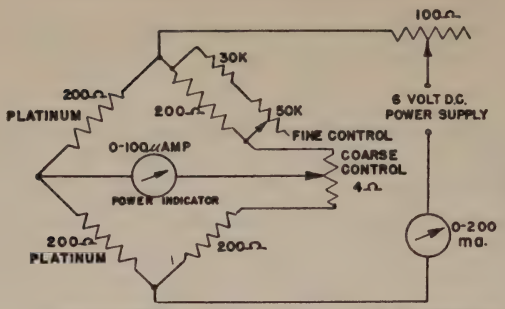


FIGURE 173.—Diagram of a Johnson power meter bridge.

d-c or a-c of low frequency through the constantan. The latter method can be used on the 3-cm. meters, but constructional details prevent its use on the 10-cm. wave guide and coaxial Johnson meters. The coaxial meters are made in just the same manner as the wave-guide meters; i. e., a thin sheet of constantan replaces the outer conductor and the platinum arm of the bridge is placed around it.

The sensitivity of the bridge is proportional to the total current passing through the bridge so that this current must be kept constant by means of a regulated power supply.

NOTE.—Care must be exercised in placing the Johnson meter in the line. The instrument, although compensated, will drift if subjected to severe thermal gradients such as are encountered near a magnetron or a sand load.

This type of meter can be placed in the transmission line and left there to monitor continuously. Another advantage over the water load is that it is suitable for use outside the laboratory. It is also insensitive to small changes in frequency. The appearance of Johnson meters is illustrated in figure 174.

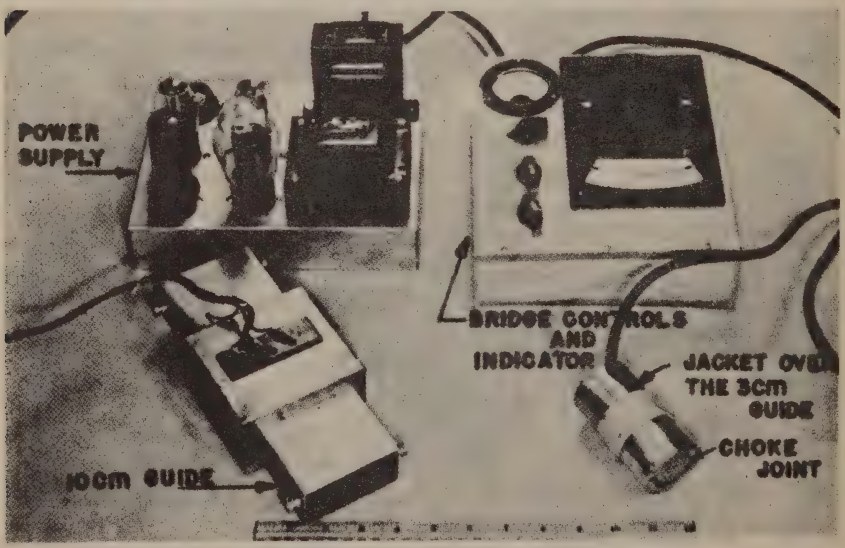


FIGURE 174.—Johnson meters for 10-cm. and 3-cm. wave guides.

(3) *Fixed attenuation plus low-power measurer.*—Other methods of measuring high power usually involve some fixed attenuation plus some low-power measuring device. For example, a probe sticking into the guide and leading to a calibrated crystal or a bolometer or thermistor bridge might be used. Care must be taken to protect the low-power devices from being burned out.

b. Low-Level Power Measurements.

General.—The most common detectors used for the measurement of low-level (microwatt) radio-frequency power are crystals, bolometers, and thermistors. The simplest to use is the crystal, since it demands no equipment other than its radio-frequency mounting and a microammeter. It is not too well adapted to accurate quantitative measurements, particularly in field test sets, *because its calibration may be changed by excessive peak power even when the average power is small.* The crystal detects by virtue of its rectifying property. The bolometer and thermistor detect through a resistance change induced by the heating effect of the absorbed radio-frequency power. Because of this dependency of resistance on power the bolometer and thermistor are usually used in bridge circuits that permit an accurate measurement of the resistance change. The detector comprises one arm of the bridge.

The bolometer usually takes the form of a 5- or 10-ma. Littelfuse whose fine platinum wire has a sensitivity of several thousand ohms per watt. A number of home-made bolometers have been prepared for use with broad-band (frequency insensitive) mountings. The bolometer's calibration is usually dependable, but it is easily broken mechanically and easily burned out with the application of too much average power. The thermistor, using a tiny bead made of a mixture of metallic oxides, has a negative temperature coefficient of resistance. Its sensitivity figure is, roughly, 10 times that of the bolometer, and it has both mechanical ruggedness and good overload characteristics. Consequently, it is the most desirable of the 3 detectors mentioned for accurate, quantitative work.

Balanced bridges.—The most precise measurements of small power are made with balanced bridges. Although the variety of balanced bridge circuits is almost without limit, the usual practice is to balance the bridge with direct-current, measuring the amount of direct-current power in the detector at balance. The r-f power is then added to the detector, unbalancing the bridge. Balance is restored by removing some direct-current power from the detector, and the change in direct-current power is used to calculate the magnitude of the r-f power added. In many bridge circuits the r-f power is exactly equal to the difference in direct-current power between these two balance conditions. Balanced bridge circuits differ principally in the way by

which they measure this direct-current power differential. It is not difficult to set up a balanced bridge that will detect a 10 microwatts of r-f power, and one can do even better with the proper equipment.

Unbalanced bridges.—The unbalanced bridge is easier and faster to use than the balanced bridge, but in most cases relies on a balanced bridge for its initial calibration. Such a bridge is balanced on direct current in the simplest case, and the deflection of the bridge galvanometer needle is noted when the r-f power is added to the detector. Previously the galvanometer scale has been calibrated in r-f power, so that the unbalanced bridge is a direct reading r-f microwattmeter. With a thermistor detector and a low resistance 0–200 microammeter it is possible to have a full-scale deflection represent 1 milliwatt, and the calibration will be approximately linear.

The thermistor is so sensitive a detector that the ultimate sensitivity of the unbalanced bridge depends principally on (1) the accuracy with which the bridge voltage supply can be regulated, and (2) the success of the thermo-stable box or compensating circuit used to protect the detector from slight variations in room temperature. A combination a-c/d-c balance circuit has been devised to guarantee the validity of the calibration regardless of the temperature at which data are taken.

Broadband mounts.—In low-level power measurement problems it is often necessary to take data over a broad wavelength band. To avoid the necessity of using tuning devices to "match in" the detector at each new wavelength setting, so-called broad-band or frequency insensitive detector mountings have been developed. These are now available for the 10-cm. band in coaxial line and the 3-cm. band in wave guide

A typical thermistor bridge.—Figure 175 shows the circuit of an a-c operated unbalanced thermistor bridge. Since a thermistor is sensitive to ambient temperature, an important problem in the design of an unbalanced bridge circuit using this type of detector is that of making the power calibration on the bridge microammeter independent of external or room temperature. In this bridge circuit that is accomplished by balancing the bridge with 60-cycle power (varying R–9), while maintaining the direct-current bridge voltage at a constant level.

The screw-driver control (R–13) is so set that the bridge is balanced, using no a-c at the highest "room" temperature at which the bridge will be expected to operate. (Extending the temperature range higher than that anticipated involves a needless sacrifice in bridge sensitivity.) At any lower temperature the bridge will not balance unless a-c is added—more a-c being needed as the temperature decreases. Since the time constant of a thermistor is such that it will

not follow a-c of frequency greater than approximately 7 c. p. s., this a-c power dissipated in the thermistor is simply a source of heat sufficient to compensate for the loss in internal heat due to a lower ambient level. The detector is, in effect, forced to operate at all times on a single temperature curve where its sensitivity at the balance resistance is constant. Finally, with a fixed, regulated d-c voltage across the bridge, the microamperes deflection per r-f microwatt will be a constant. Both the d-c and a-c voltage supplies are regulated with tungsten filament lamp bridges (the *R*–1, 2, 3, 4 and *R*–5, 6, 7, 8 meshes, respectively). The choke *L*–1 and condenser *C*–1 serve to isolate the a-c from the d-c circuits outside the detector bridge.

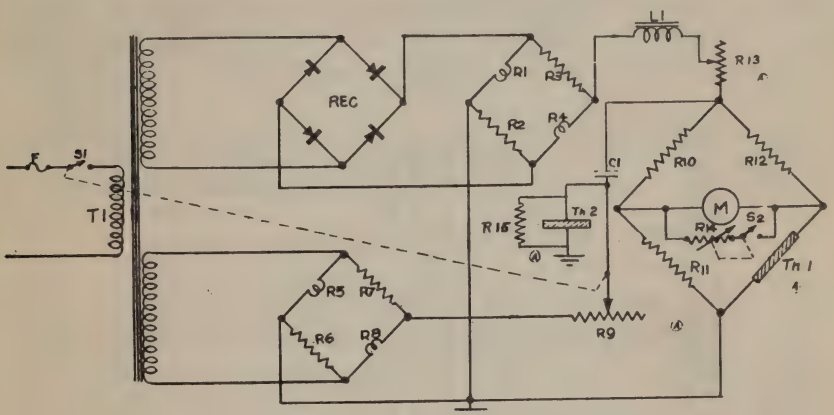


FIGURE 175.—Alternating-current operated thermistor bridge.

There are really two problems associated with temperature drifts that must be met. The use of a-c balance meets one; that is, it guarantees the validity of the calibration regardless of any given room temperature. However, it does not guarantee the validity against any changes in room temperature which occur after the bridge has been balanced and zero-set. For power monitoring purposes it is desirable to compensate the bridge against such thermal drifts so that the zero-set need not constantly be checked. This second type of temperature compensation is accomplished through the use of a "dummy" disc-type thermistor *Th*–2 and shunting resistor *R*–15, installed in the a-c supply to the detector bridge. Its use does not interfere with the calibration of the bridge, but does reduce by a factor of 50 or more the drift in zero-set due to ambient variations. The compensating thermistor is mounted in good thermal contact with the detector thermistor *Th*–1.

These unbalanced bridges are calibrated (with r-f energy) against a precision, balanced bridge.

7. TERMINATIONS

It is usually necessary to terminate a test bench set-up in a matched load so that reflections from the load will not confuse standing wave measurements, pull the r-f generator, or produce any other undesired effect.

a. Coaxial Matching Terminations.

Coaxial dry load.—A very convenient termination is the coaxial dry load which consists of 9 inches of ⅞-inch line filled with sand coated with Aquadag. The Aquadag is the only absorber, but a mixture of sand and Aquadag offers a more convenient impedance than Aquadag alone. To improve the match, the center conductor is reduced in diameter (figure 176). Although the loads can usually be made to give a standing wave ratio of 1:1 or 1:2, a double-stub tuner is ordinarily used with the load to provide a better match.

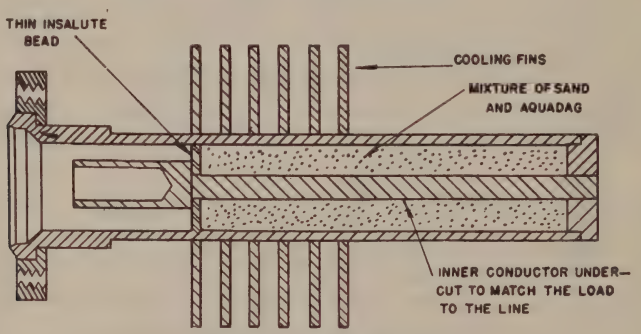


FIGURE 176.—Coaxial sand load.

In order to facilitate cooling, copper fins are soldered on the outer conductor around the first part of the absorbing mixture where most of the power is dissipated. The sand and Aquadag mixture is sealed in by means of a thin section of Insalute cement. These loads are not recommended for peak powers in excess of 50 kw. as the Insalute bead may arc over. For peak powers in excess of 50 kw. a transformation to a wave guide which is terminated in a wave-guide dry load is recommended. See figures 177, 178, and 182.

Lossy cable.—One convenient method of matching a coaxial line is to terminate it in a double-stub tuner and 6 or 10 feet of lossy flexible cable such as D163296 (Western Electric 50-ohm rubber dielectric cable) (Figure 179). Since both the incident wave and the reflected wave from the closed end are attenuated by the lossy cable, the reflected wave has small amplitude when it reaches the double-stub tuner and so can easily be tuned out. This termination is not recommended for dissipating power of the magnetron level, because the lossy line may overheat.

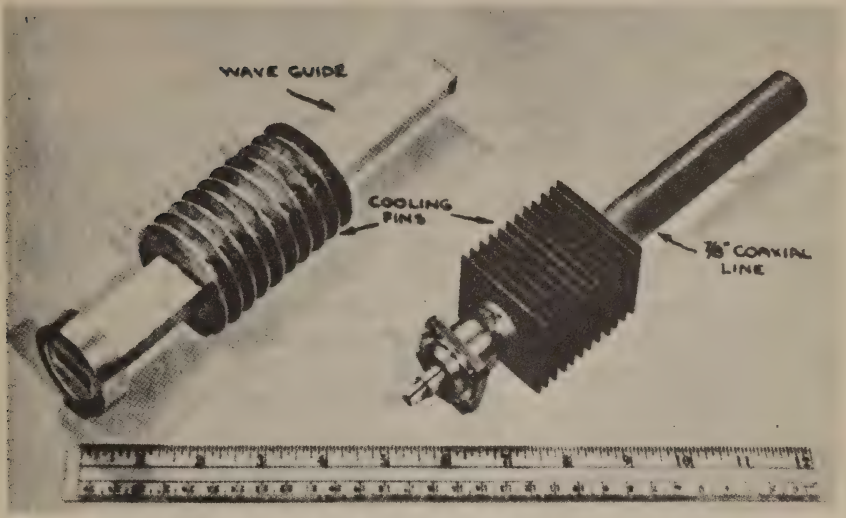


FIGURE 177.—Sand loads.

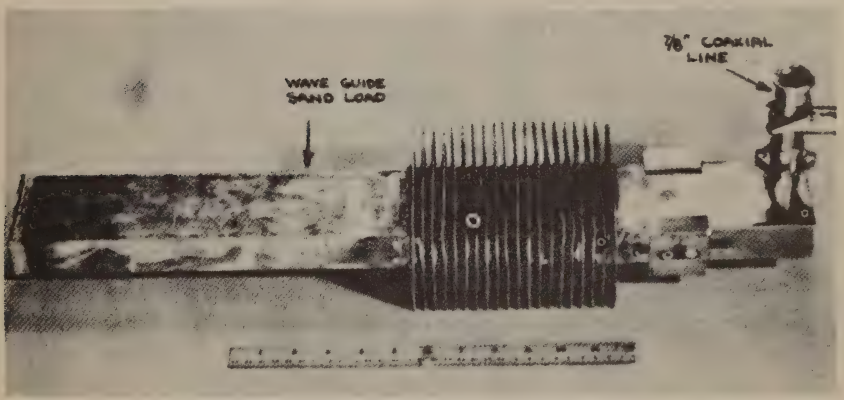


FIGURE 178.—Coaxial line to wave-guide load for high power.

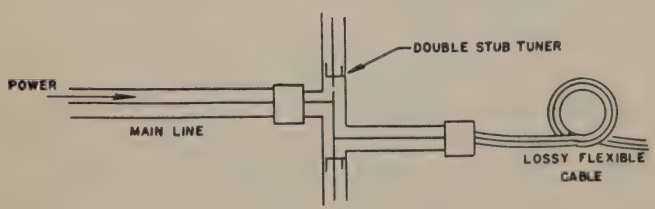


FIGURE 179.—Lossy cable termination for low-power work.

Water load.—If the bench has a water load power measurer, this device may be used as a termination. Since it is water cooled, it does not get too hot to touch as the dry loads sometimes do. For further details, see Section IV, 6a.

Antenna load.—An antenna and dish may be used as a load, but it is not recommended. The match depends so much upon reflections reaching the antenna that it is easily upset by people walking around the room.

A slightly better arrangement is an antenna inside a "black box" (figure 180). The antenna is "propeller" shaped so as not to be frequency sensitive. The inside of the box is completely covered with a "Salisbury screen" which is designed to absorb all energy incident upon it. The Salisbury screen consists of two layers: the first layer is a slightly conducting cloth whose surface resistivity is 377 ohms per square, and the second layer is a copper sheet separated a quarter wavelength from the first layer. The black box has many disadvantages and is rarely used now. It is heavy and bulky and the Salisbury screen is frequency sensitive.

FIGURE 180.—Black box load.

b. Wave Guide Matching Terminations.

The main difference between wave-guide and coaxial terminations is that of physical shape, for electrically they are much the same.

Wave guide dry load.—The wave guide dry load is made by filling a short section of wave guide with the same mixture of sand and Aquadag as is used in the coaxial dry load. It is matched to the line by a two-guide wavelength taper at the front surface. This type of match is very good. S-band dry loads have been found to produce standing wave ratios of less than 1:1 over a 20 percent band. For mechanical reasons, the front of the taper is covered by a sheet of transite or mica. The shape and plane of the taper is not important. In the circular guide shown in figure 181, the plane of taper can be

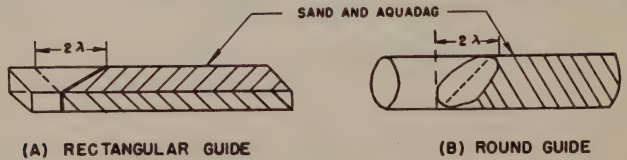


FIGURE 181.—Wave guide sand loads.

completely rotated about the guide axis without appreciably changing the match, although the guide is propagating the polarized $TE_{1,1}$ mode. Cooling fins (not shown in the figures) are usually soldered to the outside of the guide. The loads can also be water-cooled. The wave guide dry load can absorb much more power than the coaxial load without danger of breakdown and for that reason is

used for terminating high power (greater than 50 kw. peak) coaxial lines as shown in figure 182.

Resistive strip termination.—The wave guide analogue of the lossy cable termination is the resistive strip termination. It consists of a sheet of resistive material placed along the axis of the guide and in the *E*-plane. Reflections from the front end can be cancelled by means of a single screw sliding in a slot. These reflections should be minimized by tapering the resistive strip. The resistance and length of the strip should be such as to produce about 20 db. attenuation in a wave traversing it in one direction. This, like the lossy cable, is a low power level device. (See figure 183.)

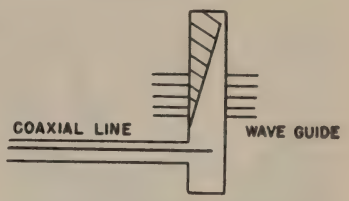


FIGURE 182.—Coaxial line to wave guide termination.

Water load.—The wave guide water load is a very satisfactory power absorber. See figure 184 and refer to the text of section IV, 6a.

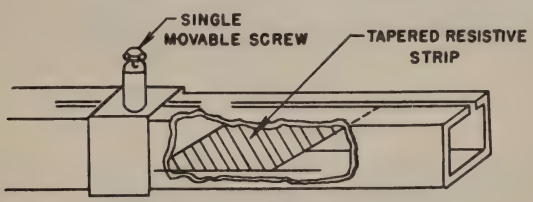


FIGURE 183.—Resistive strip termination.

Space matcher.—A wave guide can be matched to space by placing an inductive window in the guide at the proper place. This arrangement is matched at only one frequency and tuning screws must be added to match it at any other frequency. The space matcher is not used much because it is frequency sensitive and its match can be upset by reflections from people walking around the room. A space matcher and a resistor strip load are shown in figure 185.

Other X-band loads.—There are two other low-power-level terminations which are used on X-band test benches. One is a regular X-band crystal mixer. Although it has the disadvantage of being frequency sensitive, it has the advantage of being a power indicator. The other termination is the wood load which consists of a tapered wood plug cut so that it fits snugly inside the wave guide. It is usually about four wavelengths long with about a two-wave length taper at the front. It is a "stop-gap" device but a convenient one since it can be very easily made. It is not frequency sensitive and is movable inside the guide. Since it is movable, its match to the line can be roughly determined although there are reflection obstacles between it and a slotted section. For example, terminate the round guide end of a

$TE_{1,0}$–$TM_{0,1}$ transformer with a wooden plug. (See figure 186.) Since it is difficult to measure reflections in the $TM_{0,1}$ mode in the round guide, the measurement is made in the rectangular guide. It

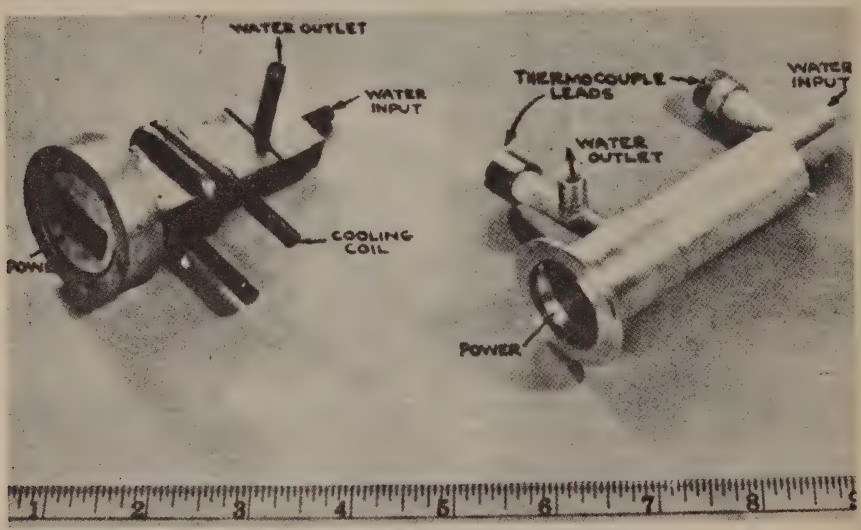


FIGURE 184.—Wave guide and coaxial line water loads.

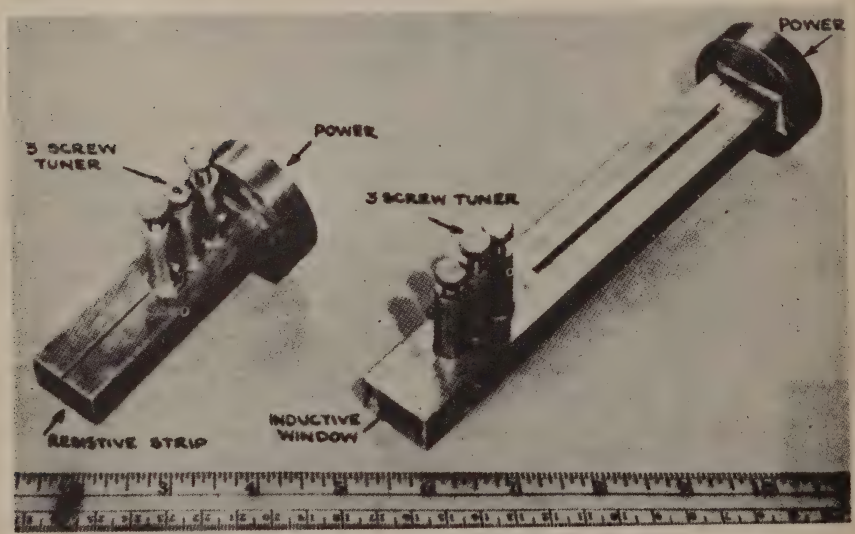


FIGURE 185.—Wave guide terminations. Left, resistive strip load; right, space matcher.

can be seen that if no change in the standing wave pattern in the rectangular guide is produced by moving the wood load, the round guide is matched.

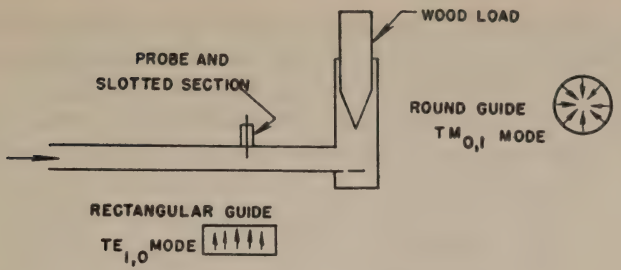


FIGURE 186.—Use of a tapered wood plug.

8. ATTENUATORS

Microwave attenuators are of two general types. The first class operates on the principle of a wave guide below cut-off. The second class operates by introducing dissipative elements to absorb energy. These will be referred to simply as the cut-off type, and the dissipative type.

a. Cut-off Attenuators.

The cut-off type is used exclusively as a variable attenuator. The calibration of such an attenuator may be computed theoretically, and may be made substantially independent of frequency. The theory underlying this type of attenuator is as follows. Propagation through a wave guide takes place without loss provided that the diameter or width of the guide is greater than the cut-off value. If these dimensions are below cut-off, then there is no longer any true wave propagation, but instead the fields are exponentially attenuated down the length of the guide. The attenuation per unit length, for any given mode, may be computed directly from theory. Furthermore, if the guide diameter is made small compared to the free space wavelength, the attenuation is then substantially independent of frequency over a wide range. The modes which are most commonly used in attenuators are the $TE_{1,1}$ and the $TM_{0,1}$. These modes and the methods used for exciting them are diagrammatically shown in figure 187. The attenuations for these two modes are as follows:

$$TE_{1,1}\text{db/cm}=\frac{32.0}{2a}\sqrt{1-\left(\frac{3.41a}{\lambda}\right)^2}$$

$$TM_{0,1}\text{ db/cm}=\frac{41.8}{2a}\sqrt{1-\left(\frac{2.61a}{\lambda}\right)^2}$$

Photographs of a magnetic attenuator for the 10 cm. band are shown in figure 188. This attenuator is designed to be used in a coaxial cable line. One loop is fixed and is not visible in the photograph. The other loop is mounted at the end of the movable tube. By telescoping the movable be inside of the wave guide below cut-off the effective length of the uide, and thereby the total attenuation, is varied.

Since the attenuation is a linear function of the separation, the dial may be calibrated directly in db. Attenuators of this type usually have an insertion loss of 10 to 20 db. at the position of maximum coupling. In the immediate vicinity of the loops, modes other than the desired one exist, and as a result the attenuation is not a linear function of the

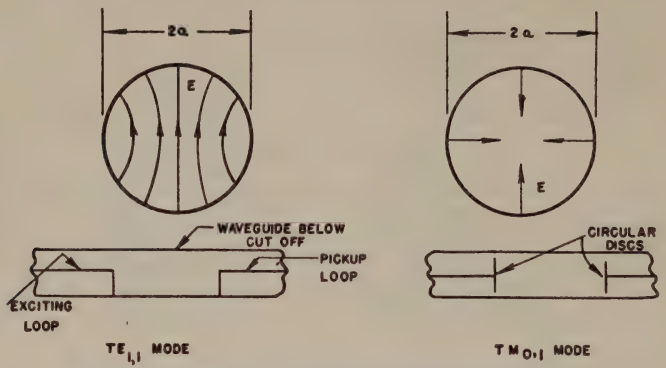


FIGURE 187.—Modes and their excitation in cut-off attenuators.

loop displacement for small displacements. The usable portion of the range therefore begins at the point where the characteristic becomes linear. This will in general occur at a total insertion loss of 25 to 35 db. for the type employing the $TE_{1,1}$ mode.

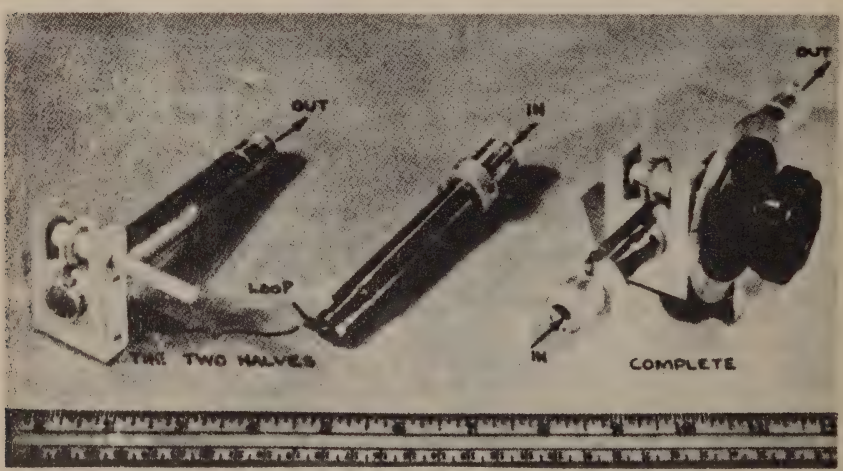


FIGURE 188.—Cut-off attenuator, magnetic type.

Since a line terminated by a loop is substantially equivalent to a short-circuited line, an attenuator of this type presents a gross impedance mismatch looking in toward either the input or output ends. This is undesirable in many applications and may be circumvented in a number of ways. The most common scheme is to pad the input and

output ends with about 10 db. of lossy cable so that reflections from the loops are substantially damped out at the far ends of the cables. Another scheme is to place a pair of resistor disks across the lines feeding the loops at the points of high impedance. The resistance of the disk should be equal to the characteristic impedance of the line, under which condition the line will be terminated properly. A third scheme which has been used with some success is to make the loop itself of resistive material and of resistance equal to characteristic impedance.

$TE_{1,1}$ attenuators are also used on the 3 cm. band. These are fed from standard rectangular guide in the manner shown in figure 189.

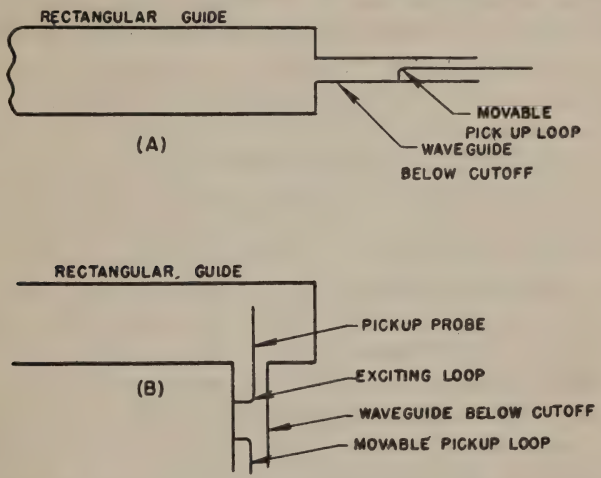


FIGURE 189.—Coupling $TE_{1,1}$ cut-off type attenuators to rectangular wave guide.

Practically all of the cut-off attenuators at the Radiation Laboratory use the $TE_{1,1}$ mode. Some Western Electric attenuators use the $TM_{0,1}$ mode. In this class of attenuators attenuation is obtained by introducing resistive elements in the wave guide or coaxial line so as to dissipate energy in the form of heat.

b. Dissipative Attenuators.

Coaxial line dissipative attenuators.—In coaxial lines the loss may be introduced either in the form of series resistance or shunt conductance. Examples of the former are the platinized glass rod and carbonized rod attenuators. Shunt conductance types are ordinary lossy cable and sand loads.

The characteristic impedance of a line which has either resistance or conductance is in general complex, and is given by

$$Z_0=\sqrt{\frac{R+j\omega L}{G+j\omega C}}$$

where R, G, L, and C are the conventional resistance, conductance, in-

ductance and capacitance per unit length of line. Z_0 is capacitive when $G=O$ and R is finite, and is inductive when $R=O$ and G is finite. Because of this fact, these attenuators present something of a mismatch when connected to a line of real characteristic impedance. This may be quite serious when R or G is made large in order to get large attentuations in a small space. This mismatch may be avoided in two ways. First, the dissipative element may be tapered so that the loss is introduced gradually. This scheme is more often used in waveguides than in coaxial lines. Second, a matching section of some sort may be used.

(*1*) *Lossy cables.*—Lossy cables are treated in section II, 1e on flexible cables and will not receive any further mention here.

(*2*) *Sand loads.*—Sand loads are attenuators used as terminating sections, to dissipate power. The space between inner and outer conductors is filled with a sand and carbon mixture which acts as the dissipative element. (See section IV, 7.)

(*3*) *Platinized glass rods.*—A platinized glass attenuator is essentially a section of coaxial line in which the inner conductor is a length of glass tubing which has been coated with an extremely thin platinum film. The film is resistive and thereby produces attenuation. By controlling the thickness of the film during manufacture various degrees of attenuation may be achieved. These attenuators have been made in both ⅝-inch and ⅞-inch standard line. The platinized glass elements are made with built-in matching sections at either end. These consist of attenuator sections, approximately $\lambda/4$ long, which have a somewhat lower film resistivity than the main attenuator section and which serve to transform the complex characteristic impedance of the main section to a real impedance equal to that of the line. This arrangement is fairly frequency insensitive. It is not difficult to obtain standing wave ratios of better than 1:1 in voltage over the 10 cm. band.

It is possible to make variable attenuators using platinized glass elements. These involve a scheme whereby either a brass tube slides over the platinized glass inner conductor, or a steel or brass rod slides inside the glass tube. That section of the platinized element which is covered by the brass tube, or in which the rod is inserted (as in the second type), contributes practically nothing to the attenuation. By sliding the tube or rod the total attenuation is varied. These variable attenuators have lower insertion losses than the more conventional cut-off type, and are reasonably well matched. The disadvantages are that they are not self-calibrating. They are somewhat large and bulky, and are difficult to make.

(*4*) *Carbonized rods.*—The carbonized rod attenuator is a series-resistance type which is used by Bell Laboratories. The center conductor is a quartz rod which has a thin deposit of carbon. There are no matching sections.

(*5*) *Polyiron.*—Another type of coaxial attenuator makes use of the r-f absorbing properties of "polyiron." This material is insulated powdered iron of the kind used extensively for magnetic cores at the lower radio frequencies. It is available either in powder form or in ready-molded cores.

If the space between the inner and outer conductors of a coaxial line is filled with polyiron powder, attenuations of the order of 5–10 db/cm at 10 cm. may be obtained. By choosing the proper diameter ratio for the line, the standing wave ratio may be kept below 1.5 in voltage over the range of 9 to 11 cm. Attenuators of this type make convenient pads. They are, however, somewhat frequency sensitive both in regard to match and attenuation, and are temperature sensitive.

Much greater attenuations may be obtained with polyiron in the molded form. Attenuations of 20 to 30 db/cm are not uncommon at 10 cm. Somewhat higher attenuations are obtained at 3 cm. Molded slugs of polyiron are extensively used in r-f line filters where it is desired to keep r-f from leaking out through power leads.

Wave guide dissipative attenuators.—Wave guide dissipative attenuators are either of the resistor-strip type, or of the sort in which the entire guide is filled with a lossy material.

(*1*) *Resistor strip.*—In the resistor strip type a strip of bakelite coated with resistance material is placed longitudinally in the guide, with the plane of the strip parallel to the short face, as shown schematically in figure 190.

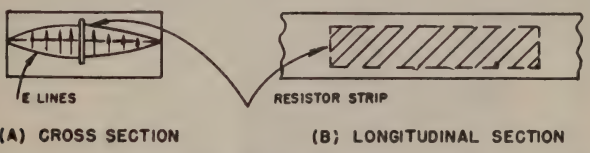


FIGURE 190.—Position of resistor strip in wave guide for maximum attenuation.

When the guide is excited in the $TE_{1,0}$ mode the electric field is parallel to the plane of the strip, currents flow in the resistive material, and energy is dissipated in the form of heat. The attenuation is a function of the surface resistivity of the strip. The lower the resistivity, the higher the attenuation (within limits) and the greater the mismatch.

Since the electric vector varies from a minimum to a maximum across the guide, the amount of attenuation with a given strip may be varied by moving the strip laterally across the guide. A variable attenuator for the 3 cm. band using this scheme is shown in figure 191 (b). Attenuations up to 70 db. have been obtained with attenuators of this type, using strips 4 inches long.

Another type of variable attenuator, using a resistor strip, is the flat type shown in figure 191 (a). In this attenuator the strip is a flap which projects through a slit in the upper face of the guide. The flap is hinged at one end so that it may be gradually lowered into the guide,

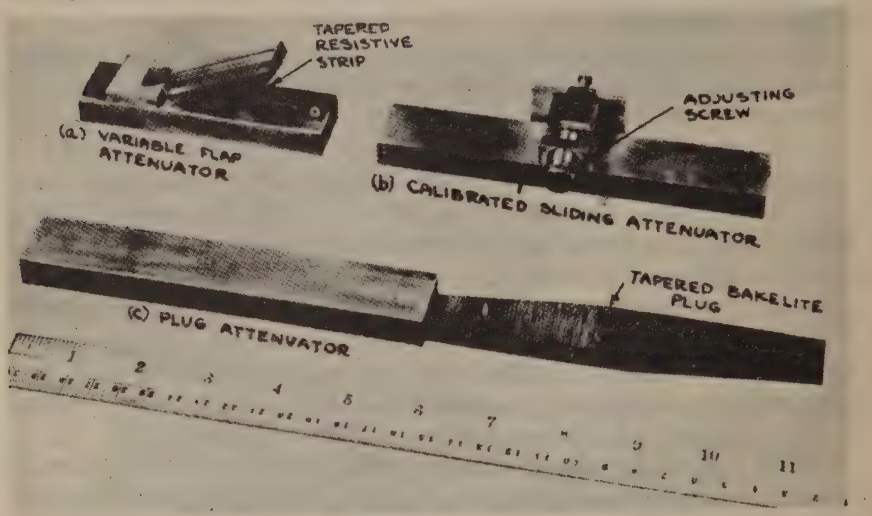


FIGURE 191.—Dissipative wave guide attenuators.

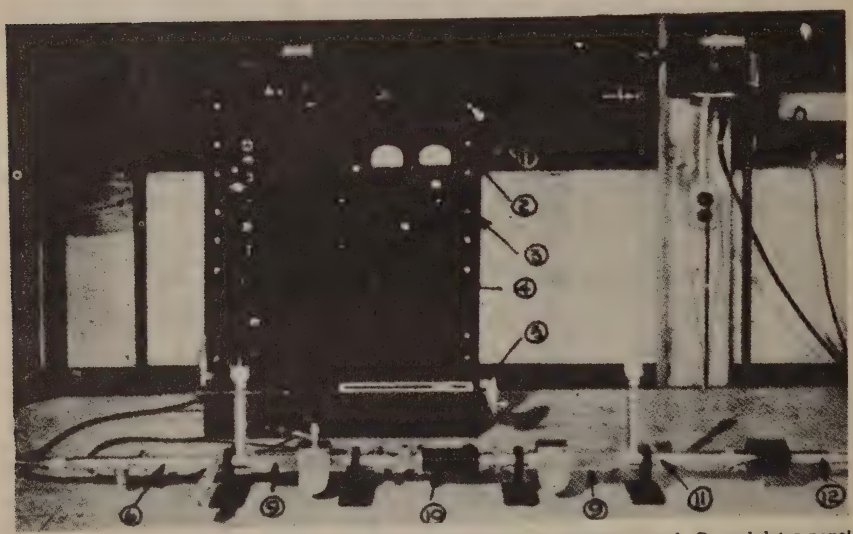


FIGURE 192.—10 cm. Klystron bench. ① power supply panel, ③ amplifier panel, ⑤ modulator panel, ⑦ Klystron panel, ⑨ wavemeter, ⑪ ⅞-inch coaxial line, ⑬ slot and probe section, ⑮ component under test, ⑰ tuner, and ⑲ load.

thereby changing the total attenuation. This type usually has a small range, generally from about zero to ten db.

Resistor strip attenuators are matched in either by mechanically tapering the strip or by cutting notches in the end of the strip. The

notched sections act as matching transformers. By making the taper long enough it is possible to get very low standing wave ratios.

(2) *Plug type.*—In the second type of wave guide dissipative attenuator, the guide is filled with conductive material; generally a plaster-of-paris and graphite mixture, the proportions of which are adjusted to give the proper attenuation. Matching is achieved either by means of a slow mechanical taper or by matching the ends. A bakelite plug has been used instead of the plaster-of-paris mixture. Attenuations of about 1 db/cm. have been reported. One such plug is shown in figure 191 (c).

Typical test bench set-ups are illustrated in figures 192 and 193.

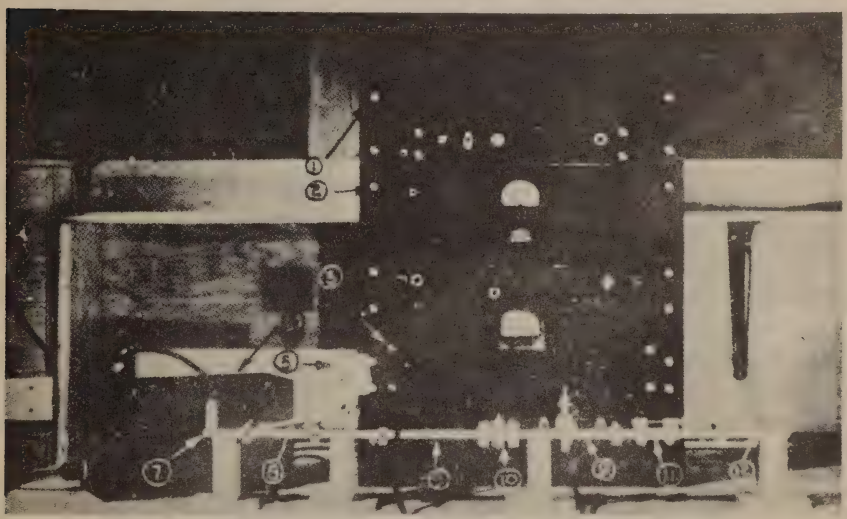


FIGURE 193.—3 cm. Klystron bench.

9. TEST EQUIPMENT, LARGER UNITS

a. R-F Spectrum Analyzer.

"Spectrum analyzer" is a name which describes only one function of a versatile r-f measuring instrument.

The elements of a spectrum analyzer are a crystal mixer (1st detector), a frequency-modulated local oscillator, a narrow-band 30 mc. i-f amplifier, r-f input attenuators, c. w. signal generators, and a cathode-ray tube indicator. The local oscillator is frequency modulated by the same sawtooth voltage that furnishes the horizontal sweep for the cathode-ray tube. The video output of the amplifier is applied to the vertical plates of the tube. A schematic diagram of the set-up is shown in figure 194.

The high sensitivity of the receiver unit permits the use of cut-off type r-f attenuators to measure large standing wave ratios, despite their high insertion loss (15 to 20 db). If a 417A Klystron or a 707A

(McNally) tube is used at 10 cm. as a c. w. signal generator, it is possible to measure standing waves of 50–60 db. The synchronized frequency-modulated local oscillator and horizontal sweep allow simultaneous observations to be made of signals 15–30 mc. apart. Thus, high selectivity (150 kc.) and large bandwidth (15–30 mc.) are obtained simultaneously, and a signal of varying frequency may be followed with only occasional retuning of the local oscillator.

If a c. w. signal is applied to the mixer, and the local oscillator adjusted so that at some point of its frequency swing the beat note between it and the signal is the same as the intermediate frequency, a

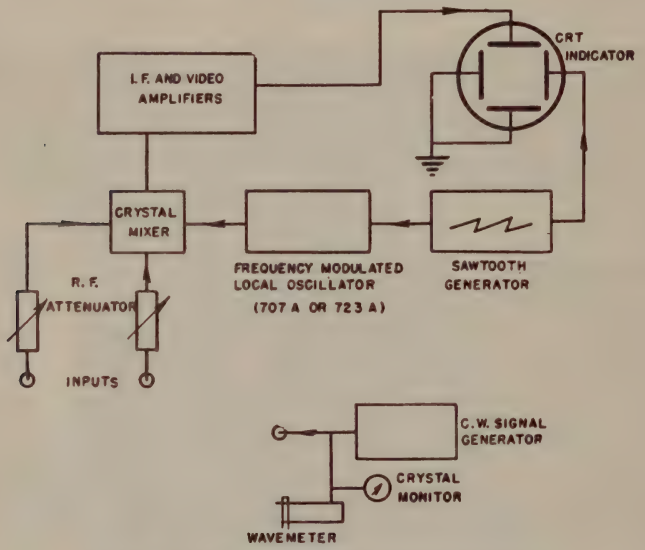


FIGURE 194.—Schematic diagram of a spectrum analyzer.

pip will appear on the cathode-ray tube. The shape of the pip will be the band-pass characteristic of the i-f amplifier (since the cathode-ray tube plots amplitude against frequency). If the signal frequency is changed, the pip will move across the screen. Several signals lying within the swing of the local oscillator may be observed simultaneously.

The spectrum analyzer was originally designed for the express purpose of examining the frequency spectra of pulsed magnetrons. The pulses of r-f energy [figure 195 (A)] may be expressed as the sum of a continuous distribution of waves such that the phases and amplitudes of these waves depends upon wavelength in a manner which describes the pulse shape.

The amplitudes of these various frequency components due to a rectangular pulse are proportional to $\sin x/x$ [figure 195 (B)], where $x=2\pi\ (f-f_0)\ T$, f_0 is the carrier frequency, and T the length of the pulse. If the spectrum analyzer is tuned to such a signal, the local

oscillator will beat with each frequency component in its range. The resultant pattern on the screen of the cathode-ray tube will have a frequency amplitude relationship proportional to $\sin^2 x/x^2$ since a crystal mixer gives a linear response in a superheterodyne and the second detector a square-law response. The width of such a pattern or distance between minima may be measured by feeding in a c. w. signal whose frequency is known or can be measured, and superimposing it on the magnetron spectrum. If the r-f pulse has frequency as well as ampli-

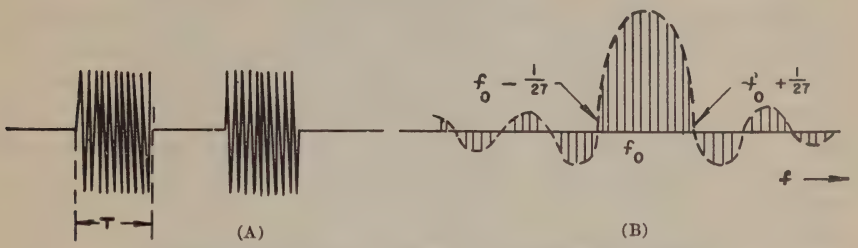


FIGURE 195.—Radio-frequency pulses and their frequency components.

tude modulation, the spectrum will be non-symmetrical about the main peak. "Double moding" or frequency-jumping shows up very plainly.

One of the main low-level uses of the spectrum analyzer is the measurement of Q's of resonant cavities or lines. These may be made either in terms of impedance or transmission vs. frequency (see section IV, 10a on measurement of Q). When high Q's are to be measured directly, it is necessary to measure changes of frequency with an accuracy equal to or better than the desired accuracy of Q measurement.

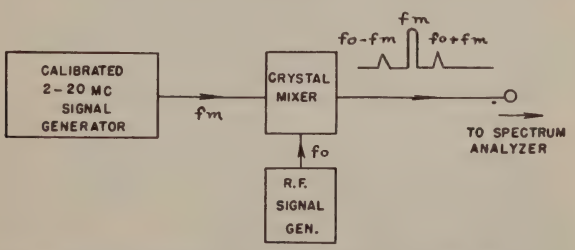


FIGURE 196.—Production of a marker signal with fine frequency control.

High-Q microwave meters have been made—some attaining a Q of 50,000. However, they are not generally available. By means of some simple auxiliary apparatus and the spectrum analyzer, frequency changes may be measured with an accuracy equivalent to a Q of 60,000 to 100,000.

The auxiliary apparatus, figure 196, consists of a variable frequency signal generator giving above 0.1 volt over a range of 4 to 30 mc., a crystal mixer and an r-f oscillator tube. The outputs of the two signal

generators are mixed in the non-linear impedance of the crystal, so that the r-f is amplitude modulated by the lower frequency signal. The output of the crystal will be made up of a carrier plus side bands and harmonics. If we limit ourselves to either the first upper or lower sideband, we see that its distance from the carrier is just equal to the modulating frequency. The r-f oscillator tube is adjusted so that the carrier falls just outside the range of the frequency-modulated local oscillator of the analyzer, and the modulating frequency adjusted till the first side band appears on the screen. This may now be used as a frequency marker, and superimposed on any unknown frequency. The difference between two frequencies may be read directly from the modulating signal generator's dial (which may usually be read to 10–20 kc.). Accurate superpositions of two frequencies is made possible by the appearance of a "rain" or beat note when the two are exactly equal (figure 197). The rain is due to the beating of the two identical signals in the mixer. It is possible to bring two signals together on the cathode-ray indicator if the frequency of one is 30 mc. below that of the local oscillator at that point and the frequency of the other is 30 mc. above, but in this case no rain will appear.

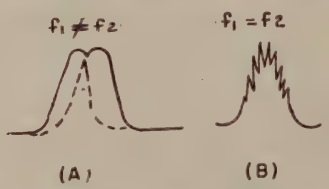


FIGURE 197.—Appearance of pattern when two signals are superimposed in a spectrum analyzer. (A), nearly superimposed; (B), exactly superimposed.

A simplified version of the above scheme of obtaining a fine control marker signal consists of a 100 mc. oscillator heavily modulated by 1 mc. This produces both amplitude and frequency modulation producing an infinite line spectrum about the carrier with line separation of 1 mc. These frequencies are fed into a crystal which generates harmonics of all the frequencies and also produces cross modulation making it possible at 3,000 mc. to find a harmonic spectrum with 1 mc. separation. These appear on the analyzer as a series of pips 1 mc. apart that can be used as a frequency increment scale on the analyzer.

Figure 198 shows a spectrum analyzer set up for measuring the cold inpedance of a magnetron

b. R-F Envelope Indicator.

Use.—This instrument (figure 199) is designed to show accurately on an oscilloscope screen the shape of the envelope of the r-f pulse radiated by a magnetron or other u-h-f generator. The presence of double modes (i. e., the magnetron's oscillating during some pulses

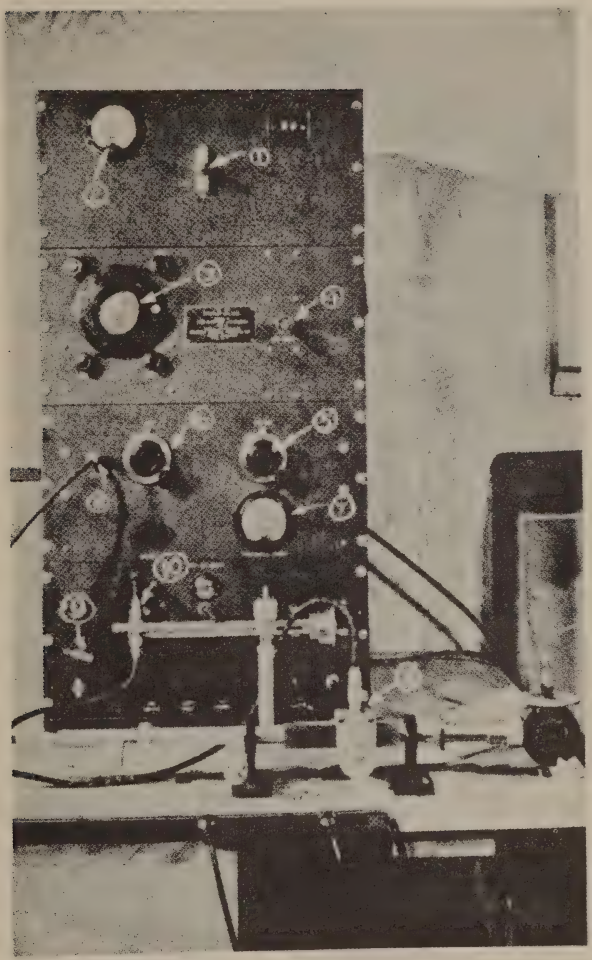


FIGURE 198.—10 cm. spectrum analyzer set up for measuring the cold impedance of a magnetron. ① control for c. w. signal generator (a 707A in this case), ② meter for monitoring the signal generator which reads the current from the crystal at ⑩, ③ cathode-ray tube indicator, ④ control for the f-m 707A, ⑤ and ⑥ attenuator controls for the two inputs, ⑦ meter reading from crystal monitoring the f-m 707A, ⑧ input #1, ⑨ input #1 (not used in this picture), ⑩ junction in the output of the signal generator which feeds the coaxial wavemeter, monitoring crystal, and the load, ⑪ slotted section and probe, which feeds into input #1, and ⑫ magnetron under test.

FIGURE 199.—Radio-frequency envelope indicator.

in one mode and during other pulses in another mode) is evident at a glance by double traces on the screen. The relative brightness of the two traces shows the relative fraction of the pulses spent in each mode. A change in mode during each pulse is also evident by a break in the top of the r-f pulse.

The indicator can be particularly useful in tuning up systems for maximum output with a clean pulse (i. e. one with a single frequency).

This instrument is also used in TR box testing. The "spike" and "flat" of the envelope indicate the power leaking through the TR box before and after it starts firing during the transmitted pulse

General description.—The various parts of an r-f envelope indicator are schematically shown in figure 200. An r-f signal fed to the

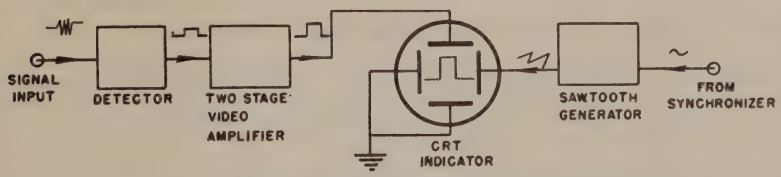


FIGURE 200.—Schematic diagram of a radio-frequency envelope indicator.

indicator is detected by a 954 acorn tube and amplified by a two-stage, wide-band video amplifier. It is then applied to a 5-inch cathode-ray tube which has a very fast sweep. The sweep is usually synchronized by means of a sine wave from the synchronizer. At 10-cm. the signal is usually fed into the indicator through a flexible cable. In the 3-cm. region rigid wave guide connections are usually used. However, the rigid plumbing can be avoided by using a crystal converter at the pick-up point and then feeding the video signal to the indicator through a flexible cable. In this case the acorn tube acts as an amplifier, and additional sensitivity can be gained in this manner. However, this arrangement has the disadvantages usually encountered with crystals, i. e., burn-out, saturation, change of calibration, etc.

R-f envelope indicators are now being combined with synchroscopes by installing the receiver strip inside the regular synchroscope case. In this manner it is possible to view both the pulse to the magnetron and the pulse envelope from the magnetron on the same screen by throwing a switch.

c. Test Sets.

Test sets are designed for use in checking systems in the field and for bench testing of plumbing. They provide means for monitoring the the power output of the transmi ter, the receiver sensitivity, the band-pass of the receiver, and the general functioning of the indicator circuit. The 10-cm. and 3-cm. sets are much alike but they, of

course, differ in the type of oscillator and the details of the r-f plumbing. The Type A (10 cm.) test set will be described as typical.

Type A test set.—The type A test set is shown schematically in figure 201.

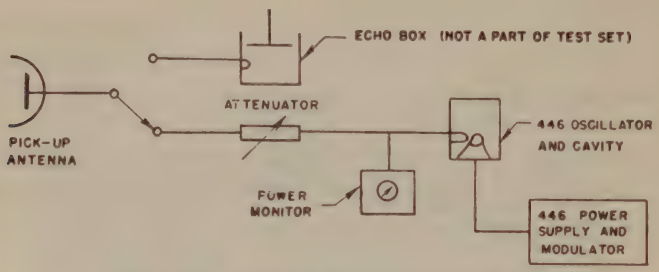


FIGURE 201.—Schematic diagram of type A test sets.

To test the power output of the radar system the energy from the pick-up dipole is fed through the variable attenuator and into the power monitor, an unbalanced thermistor bridge. The pick-up dipole is permanently mounted on the radar installation and also may be used with an echo box for rough overall system checks. The resonant cavity also serves as the cavity for the 446 oscillator (Lighthouse tube) when the test set is used as a signal generator. The thermistor bridge is much like the one described in section IV, 6b. It is an unbalanced, calibrated bridge with a single temperature compensating circuit; that is, the bridge can be zero adjusted by means of an a-c heating circuit so that the calibration holds at any given room temperature, but this compensation does not guard against thermal drifts which occur after the bridge has been zero set. The cavity is adjusted to resonance as indicated by a dip on the bridge indicator. The attenuator between the dipole and the cavity can be adjusted to give a reference reading on the bridge indicator. The tuning procedure causes the cavity to be tuned to the correct wavelength so that when the oscillator is turned on, it is automatically tuned to the frequency of the transmitter.

After the power output of the transmitter has been checked, the Lighthouse tube is turned on. The power output of this oscillator is also measured on the thermistor bridge which was used in measuring the output of the transmitter. The 446 oscillator is modulated with a variable pulse. It is variable in width from 0.5 to 5 microseconds and variable in phase from 5 to 150 microseconds after the applied trigger pulse. This r-f pulse is fed through the pick-up antenna and into the radar receiving set, where the pulse can be viewed on the indicators. The attenuator in the test set is adjusted until the pulses are just visible. The reading of the attenuator is then a criterion of the signal-to-noise ratio of the radar receiver.

The test set is available with any one of the three sizes of cavities, each covering a range of about 6 millimeters (180 mc) centered on the S_A, S_S, and S_G bands. The set can be synchronized with the radar system by means of either a positive or negative trigger, and any repetition rate between 500 and 2,000 cycles per second can be used. The set can also supply a 1,500-cycle trigger.

The test set, not including the pick-up dipole and connecting cable, is contained in a metal cabinet measuring 13 inches wide by 8 inches high and 9 inches deep. Its weight is approximately 35 pounds, and the power consumption is about 70 watts.

10. MEASUREMENTS

a. *Q* Measurements.

Transmission method —The determination of the Q of a resonant device such as a TR box or some other variety of cavity resonator involves measurement of the resonant frequency f_0 and the frequency difference Δf between the half-power points of the resonance curve. We then have $Q=f_0/\Delta f$.

If the power can be fed through the device, an arrangement as in figure 202 may be used to determine Q.

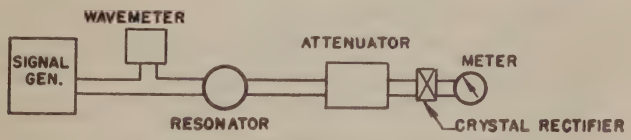


FIGURE 202.—Measurement of Q, transmission method.

With this arrangement, providing a good match looking back into the generator exists, resonance is indicated when the output meter shows a maximum. The resonant frequency may then be read off the wavemeter. The next step is to reduce the attenuation 3 db and adjust the frequency of the signal generator until the meter reading drops to the value at resonance. This is done for frequencies above and below resonance and $Q=f_0/\Delta f$ is then calculated.

Several precautions should be taken: Sufficient lossy cable should be used in order to decouple the components from one another and to minimize the effects of reflections from connectors, etc.; the output of the signal generator should be constant over the frequency range involved; and it is probably a good idea to have the wavemeter set off resonance at all times except when measuring the frequency.

For high Q values it is necessary to make a more accurate measurement of the frequency difference than is possible with an ordinary signal generator.

For this purpose it is convenient to use a spectrum analyzer provided with an auxiliary fine control marker signal which is obtained by

mixing a low-frequency (4–30 mc.) oscillator with an ultra-high-frequency oscillator (such as a Model 417 Klystron, if working at 10 cm.). The production of this fine control marker signal is discussed in section IV, 9a. The spectrum analyzer set-up for measuring the Q of a high-Q cavity by the transmission method is shown below in figure 203.

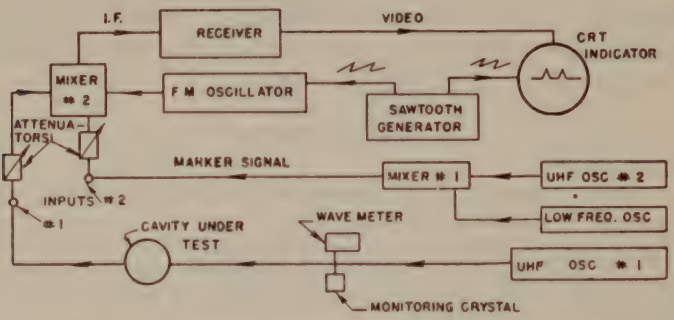


FIGURE 203.—Schematic set-up for measuring high-Q cavities.

The output of the c. w. oscillator #1 is fed through the cavity whose Q is being measured and into input #1 of the analyzer. The marker signal is fed into input #2. The actual procedure followed in measuring high-Q cavities may be outlined as follows:

(*1*) Find the resonant frequency of the device under test and locate a pip from oscillator #1 on the screen of the cathode-ray tube indicator. Measure the resonant wavelength (λ_0) by means of the wave meter.

(*2*) With the low-frequency oscillator set at some convenient value (6–10 mc.) adjust oscillator #2 so that the first side band is on the screen, and the carrier is just off the screen (on either the right or left). Then leave oscillator #2 set for all the remaining steps.

(*3*) Adjust the frequency of the low-frequency oscillator until the first side band coincides with the signal from oscillator #1. This first side band is the marker signal. Note the frequency (f_1) of the low-frequency oscillator, and then shift the marker signal to one side by adjusting the low frequency oscillator.

(*4*) Note the height (h_r) on the screen of the signal from oscillator #1 at resonance. Reduce attenuation in input #1 by 3 db. Then adjust oscillator #1 to the frequency above resonance at which the height of the signal drops to the height h_r.

(*5*) Adjust the low-frequency oscillator until the marker pip coincides with that from oscillator #1. Note the frequency f_2 of the low-frequency oscillator.

(*6*) Adjust oscillator #1 to the frequency below resonance at which

its signal height drops to h_r. Repeat step 5, this time reading the frequency f_3.

(7) Calculate Q from

$$Q=\frac{f_0}{\Delta f}=\frac{c}{\lambda_0}\frac{1}{(f_2-f_3)}$$

PRECAUTIONS.—The outputs of both oscillator #1 and the frequency modulated oscillator should be constant over the range considered. Coincidence of the marker and main signal may be judged by the appearance of "rain" caused by the beating of the two signals. When this appears, the usual resonance curve or pip on the scope will appear as the envelope of the beats. It may be convenient to calibrate the scope screen in case the attenuator must be used in its nonlinear and uncalibrated region. This can be done by adjusting the base line and height of the signal to a given reference value and then introducing 3 db of attenuation in the signal input and noting the corresponding height of the pip.

Then in step 4 above, having set the main signal to the reference value, adjust the frequency of the main oscillator until the height falls 3 db. Be sure that the wave meter is tuned off resonance while making the measurements.

These measurements give the "loaded" Q which includes the effects of the coupling into the device being measured. The procedure to determine the unloaded Q is as follows: Set the device at resonance and measure the transmitted power (P_0). Then remove the device from the line and replace it with a reflectionless and lossless connector and measure the input power (P_i). The unloaded Q_0 is then computed from the loaded Q_L and the transmission loss: $\frac{Q_0}{Q_L}=\frac{2}{1-T}$ (T is *not* expressed in db in this expression but is the ratio P_o/P_i). The expression assumes a matched input to the device and is applicable to most cases in practice.

One input method.—Measurements of tunable low-Q devices which are loaded with only one input (output shorted) may also be measured with the spectrum analyzer. In this case the apparatus is set up with the main signal fed through a slotted line section into the tunable resonant device. A marker-oscillator and low-frequency oscillator may be used as before. The steps in the measurement are:

(1) Set main oscillator to wave length at which Q is desired.

(2) With the device detuned as far as possible find the position of the minimum in the standing wave in the line between oscillator and object.

(3) Set the probe of standing wave detector $\lambda/4$ from position found in (2). (This places it at voltage maximum.)

(4) Tune the device so that the probe voltage now becomes a minimum in position (3). This means that the device is now tuned to resonance. (The tuning should not be changed during the remaining steps unless it is found that the resonant frequency differs from the desired frequency—see (6).)

(5) Measure the standing wave ratio r_0 at resonance.

(6) Using the marker pip, measure the standing wave ratio at 1 or ½ mc. intervals and plot a curve of standing wave ratio against frequency. If the resonant frequency as indicated by the minimum of the curve does not agree with the desired frequency, change the tuning slightly and repeat the curve. However, unless great accuracy is desired, the resonant frequency may differ slightly from the desired frequency without seriously affecting the value of Q.

(7) Using the value of the standing wave ratio at resonance (r_0) compute

$$r_{1,2}=\frac{r_0+1+\sqrt{r_0^2+1}}{r_0+1-\sqrt{r_0^2+1}}$$

and find the points of intersection of this value and the curve plotted in (6). This gives Δf.

(8) $Q_L=\frac{f_0}{\Delta f}$ where Q_L is the loaded Q.

(9) To determine the unloaded Q it is necessary to distinguish two cases arising, in part, from the characteristics of the losses in the device. Introducing a factor β which may be either greater or less than unity (for most cases $\beta>1$), the unloaded Q is given by

$$Q_0=(1+\beta)Q_L$$

where β is determined from the standing wave ratio at resonance (r_0) and the behavior of the position of the minima. For $\beta\ (=r_0)>1$ the minima will shift by $\lambda/4$ as the wave length goes off resonance sufficiently far, while for $\beta\ (=1/r_0)<1$ the position of the minima far off resonance will be the same as at resonance, as shown in figure 204.

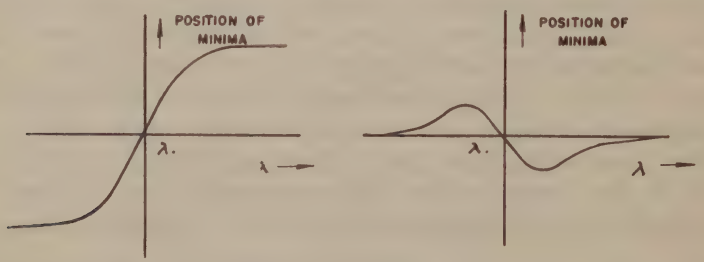


FIGURE 204.—Phase shift of the minima for $\beta>1$ and for $\beta<1$.

b. F-M Bench Measurements.

The wide range of electronic tuning available with a 723A/B oscillator allows the tube to be employed in a frequency-modulated system for certain radio frequency measurements. Where it is applicable the system greatly facilitates measurements and is particularly useful for production testing. A block diagram of the system employed is shown in figure 205.

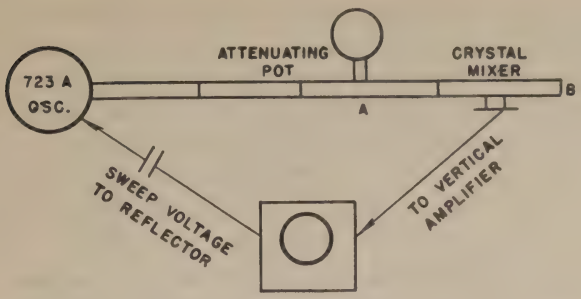


FIGURE 205.—Frequency-modulated 3-cm. band test system.

The saw tooth sweep voltage which is normally applied to the horizontal plates of the cathode-ray tube is connected in addition to the 723A/B. Then as the voltage on the reflector electrode is changed from its direct-current potential by the saw tooth sweep the tube will begin oscillating at a given voltage, reach a maximum, and finally stop oscillating as the sweep voltage increases. The pattern seen on the oscilloscope is shown in figure 206. The oscillations start at point *X*,

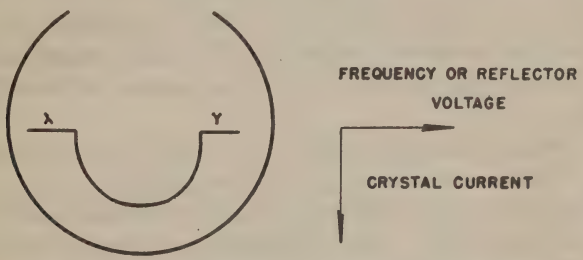


FIGURE 206.—Oscillograph pattern—frequency-modulation system.

continue with varying intensity to point *Y* where they stop. Except at points close to the edges *X* and *Y* the frequency changes almost linearly 50 to 100 megacycles over the pattern. If the oscillator is sufficiently isolated from its load by an attenuator, the frequency is determined only by the electrode voltages, and there is a one to one correspondence between frequency and a point on the oscillograph screen. This system may be used either as an absorption or a transmission measuring system.

Transmission measurements.—If a resonant cavity such as a TR box is placed in series with the line at point *A* (figure 205), and if the mixer plunger at *B* is replaced with a cavity-type wavemeter, the pattern on the oscillograph will look like figure 207.

The pip at 3 is caused by the wavemeter resonance and it can be moved along the pattern by tuning the wavemeter. By a change in the frequency of the oscillator the whole pattern can be moved along in frequency, that is, shifted to the right or left on the screen.

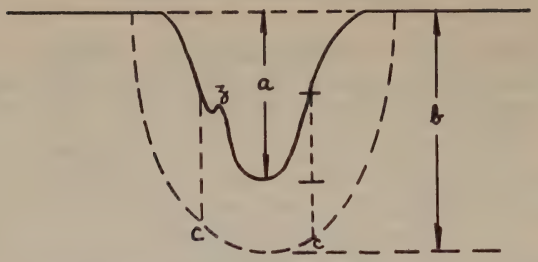


FIGURE 207.—Change in oscilloscope pattern caused by resonant cavity.

We can determine following quantities from this picture:

(*1*) The resonant frequency of the cavity, by tuning the wavemeter until the pip falls on the peak of the pattern when the oscillator is tuned for maximum transmission through the cavity.

(*2*) The loaded Q of the cavity by measuring the frequency breadth of the pattern between points where the transmission has fallen to half its maximum value. Since the crystal output is proportional to power this would be approximately half way on the pattern. If it is desired to make accurate measurements, the half power point could be found by means of an attenuator. Also allowances must be made for the difference between the points C and the center of the curve when the cavity is not in the circuit. In general this method is used for $Q_L > 300$.

(*3*) The transmission by taking the ratio of a/b where b is the height of the curve when the cavity is not present.

By placing a calibrated wavemeter and an uncalibrated one on the system at the same time, the latter can be calibrated by bringing its pip in coincidence with that of the calibrated wavemeter.

(*4*) *Reaction-type Measurements.*—If we place a resonant cavity at B in place of the plunger and a wavemeter at A on a side stub (figure 205), we can determine a number of properties of the cavity. The oscillograph pattern will look as shown in figure 208.

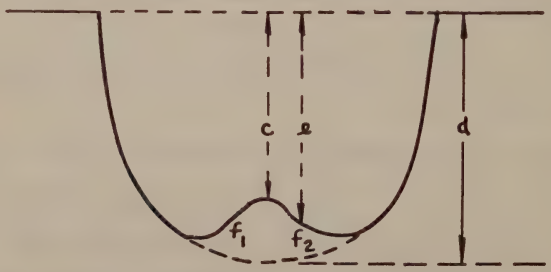


FIGURE 208.—Pattern using frequency modulation system for absorption measurements.

If we adjust the distance from the crystal to the input window so that the pattern is maximum when the cavity is not resonant, the

following things will occur at resonance: Some of the energy will go into the cavity and be absorbed, and the cavity will no longer appear as a short circuit at the window; the crystal and cavity will no longer be matched to the line; and less power will be recorded by the crystal.

The reaction method can be used to measure cavity conductance, window susceptance, and cavity Q.

Referring to figure 208:

$$\frac{c}{d}=\frac{P_r}{P_0}$$

where P_r=power at resonance
P_0=power far from resonance

From this we can calculate cavity conductance

$$\frac{c}{d}=\frac{P_r}{P_0}=\frac{4G^2}{4G^2+4G+1}$$

which can be solved for G and using

$$\frac{P\frac{1}{2}\text{ power}}{P_0}=\frac{8G^2}{8G^2+4G+1}$$

which is equal to the ratio e/d. This determines two points f_1 and f_2. The frequency difference between these two points give the loaded Q from

$$Q_L=f/\Delta f$$

and from G and Q_L we get

$$Q_0=\frac{G+1}{G}Q_L$$

and the cavity shunt resistance $R_s=Q_L/G$

c. Loss Measurements.

Insertion loss method.—The insertion loss method of measuring losses in lines may be easily applied if a calibrated r-f attenuator is at hand. The procedure is to feed power into the piece to be measured then through an attenuator to an output indicator—either a meter or a spectrum analyzer screen. The output is noted, the piece then removed, and the attenuator cranked out until the output is the same as before. The difference in attenuation is then the desired insertion loss. In using this method it is possible that reflections will cause considerable difficulty and the usual practice is to provide at least 10 db of attenuation on each side of the piece being measured.

Two-probe power-reversal method.—Another method which has been very successful is the two-probe power-reversal method. This has been used to measure losses ranging from 5 percent to 40 percent (0.2 to 2 db) with an accuracy of the order of 2 percent to 5 percent.

The apparatus is arranged for a typical 10 cm. bench as shown in figure 209.

With cable A connected to the generator, and cable B free, tuner B is adjusted until probe B indicates no standing wave. Next cable A is disconnected and the generator connected to cable B so the power is going through in the opposite direction, and tuner A is adjusted for no standing waves as indicated by probe A. This should be done

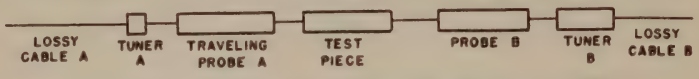


FIGURE 209.—Two probe power reversal set-up at 10 cm.

before taking any measurements. Obviously, the standing waves introduced by the test piece may be obtained for either direction of power flow, and if the position of a voltage minimum is noted, the impedance at any desired point may also be found.

The current in either probe will be proportional to some "power" n of the voltage in the line, the proportionality factor S depending on the probe insertion, detector sensitivity and tuning and amplifier gain, while the power depends on the "law" of the detector and amplifier as a whole. Thus let

$$\text{current from probe } A = I_A = S_A V_A^n$$
$$B = I_B = S_B V_B^n$$

Then the power in the line is:

$$P_A = \frac{V_A^2}{Z_0} = (C_A I_A)^{2/n} \text{ where } C_A = \frac{1}{S_A Z_0^{n/2}}$$

Then for case I (power into A):

$$T_1 = \left(\frac{\text{Power out}}{\text{Power in}}\right)_{\text{I}} = \frac{P_B'}{P_A'} = \left(\frac{C_B I_B'}{C_A I_A'}\right)^{2/n}$$

the prime denoting case I values.

For case II (power into B):

$$T_2 = \left(\frac{\text{Power out}}{\text{Power in}}\right)_{\text{II}} = \frac{P_A''}{P_B''} = \left(\frac{C_A I_A''}{C_B I_B''}\right)^{2/n}$$

Multiplying these expressions together we see that the reversal of power causes the proportionality factors to enter into the expressions in such a way that they cancel out:

$$T^2_{\text{mean}} = T_1 T_2 = \left(\frac{C_B I_B'}{C_A I_A'}\right)^{2/n} \left(\frac{C_A I_A''}{C_B I_B''}\right)^{2/n} = \left(\frac{I_B' I_A''}{I_A' I_B''}\right)^{2/n}$$

$$T_{\text{mean}} = \left(\frac{I_B' I_A''}{I_A' I_B''}\right)^{1/n}$$

So far we have assumed that no standing waves exist in the slotted section. When standing waves exist we take:

$$I_A' = [I_A'(\text{max.}) \cdot I_A'(\text{min.})]^{\frac{1}{2}} \text{ etc.}$$

Then

$$T_{\text{mean}} = \left(\frac{I_B'(\text{max.}) I_B'(\text{min.})}{I_A'(\text{max.}) I_A'(\text{min.})} \cdot \frac{I_A''(\text{max.}) I_A''(\text{min.})}{I_B''(\text{max.}) I_B''(\text{min.})} \right)^{1/2n}$$

In case some "slope" exists in the standing wave detectors (this shows up when comparing successive maxima and minima) the practice is to take three (*not* four) readings with the probe (i. e. max. min. max. or min. max. min.) and average the outer pair to get a number corrected for slope so as to be comparable with the center reading. This same procedure is followed when determining the standing wave ratio. All subsequent readings should be taken in the same region of the slot when the power is reversed in order to get comparable readings.

The determination of the value n should be done with care. Bolometers are quite close to square law ($n=2$) but it may happen that the amplifiers used in connection with the bolometer are not linear, and it should be noted that n is the *over-all* law of the detector system. Crystals of course are known to differ from square law and from crystal to crystal.

Attenuation in cables.—The method discussed in the preceding paragraphs is the usual method. Another method by which the attenuation in cables or other coaxial lines is measured is described in a Sperry report by W. W. Hansen on the "Measurement of Coaxial Line Loss." In this method the line to be measured, which must be short, is terminated by a short short-circuited lossless line of variable length (e. g. coaxial shorting plunger) and is connected on the other line to standing wave detector. (See figure 210.) If there

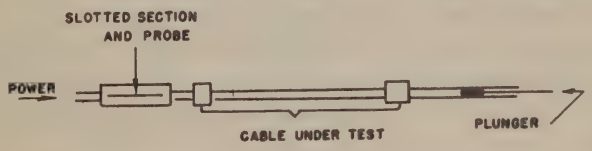


FIGURE 210.—Measuring the attenuation constant of a cable.

is a perfect impedance match between the line to be measured and the standing wave device, the standing wave ratio will be idependent of the variable line length, and we will have*

$$\tanh \alpha l = \frac{V_{\text{min}}}{V_{\text{max}}} = \frac{1}{r}$$

*From a paper on "Measurement of Coaxial Line Loss", by W. W. Hansen, Sperry Gyroscope Co., August 15, 1941.

where α is the attenuation constant and l is the length of the line being measured. If αl is small, the standing wave ratio may be very large so that it may be more convenient to determine x_0, in which case

$$\alpha l = \frac{2\pi x_0}{\lambda} = \frac{1}{r}, \quad \alpha l \ll 1$$

where x_0 is the distance from the minimum V^2 to a position of twice this V^2.

More commonly the standing wave ratio depends on the line length so one determines the maximum and minimum standing wave ratios (r_1 and r_2) as the length of the shorted line is varied and we have

$$\tanh \alpha l = \sqrt{\frac{r_2}{r_1}}$$

if the minima for r_1 and r_2 are in the same places, and if the minima are in different places

$$\tanh \alpha l = \frac{1}{\sqrt{r_1 r_2}}$$

To express the cable loss in db/cm:

$$\text{db/cm} = 8.7\alpha$$

d. Impedance Measurements.

The standing wave ratio and the phase of the standing wave pattern relative to any point along the line uniquely determin the normalized impedence, or admittance, at that point. The actual determination of the impedance from these two data is most conveniently accomplished by means of some form of impedance chart, which shows how impedances transform along a line. The chart most commonly used in the Radiation Laboratory is the circular Smith chart. (See Appendix, section 3.)

As an example, consider the use of the chart in determining a terminating impedance of a coaxial transmission line. A slotted section is placed before the termination, and the standing wave ratio (r) and the position of a minimum point (voltage node) are determined by means of a probe and an indicating device as shown in figure 211.

Suppose now that at a maximum the indicator reads 0.990 ma.

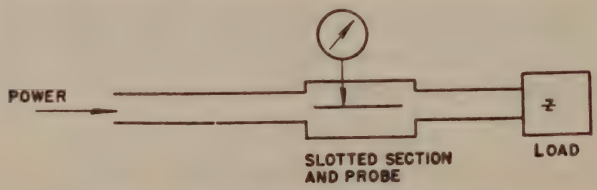


FIGURE 211.—Measuring the load impedance.

and at a minimum 0.110 ma. and that we know our probe and indicator device as a whole to be perfectly square law. Then the standing wave ratio in voltage is:

$$r=\left(\frac{\text{max.}}{\text{min.}}\right)^{1/n}=\left(\frac{.990}{.110}\right)^{1/2}=(9.00)^{1/2}=3.0$$

This fact tells us that the load impedance is some place on the "$r=3$" circle of the impedance chart (The "$r=3$" circle is tangent to the "resistance=3" and the "resistance $(=1/3)=0.33$" circles of figure 212.)

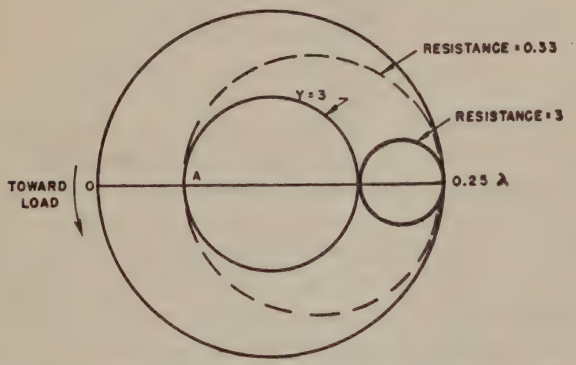


FIGURE 212.—Location of the "$r=3$" circle.

Now let us further suppose that a minimum point in the standing wave pattern is 1.9λ away from the load. The minimum point locates us at A in figure 212 (resistance is a minimum). Moving down the line toward the load moves us counter-clockwise around the impedance chart on the "$r=3$" circle. Moving a half wavelength along the line is equivalent to moving once completely around the "$r=3$" circle. Therefore moving 1.9 wavelengths down the line moves us 3.8 times around the chart. The periphery of the chart is marked in units of λ for convenience. Moving 1.9λ from the minimum brings us to point B (see figure 213) which is the intersection of the "$r=3$" circle and the "0.40λ radius arm." Point B is also the intersection of the "resistance=0.48" and the "reactance$=+\ j\ 0.61$" circles. Therefore, the normalized load impedance is:

$$z=0.48+j\ 0.61$$

The impedance chart can also be used as an admittance chart, in which case the constant resistance circles become constant conductance circles and the constant reactance circles become constant susceptance circles. If the admittance y is defined

$$y=g+jb \text{ (Instead of } y=g-jb)$$

then an inductance will be a negative susceptance and a capacity a positive susceptance and there is no other change in the use of the

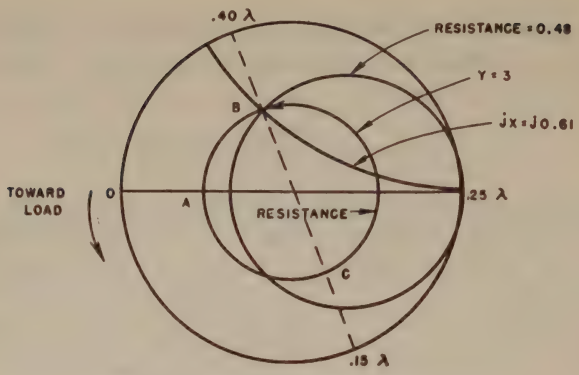


FIGURE 213.—Determination of the load impedance.

chart. The directions of rotation in moving toward the load and generator remain the same, and positive reactance and positive susceptance lie on the same half of the chart. One thing to note is that when the chart is to be used as an admittance chart, minima of the standing wave patterns occur on the right half of the real axis (conductance greater than 1.0) whereas in the impedance case they occur for equivalent reasons on the left-hand side (resistance less than 1.0).

The circular impedance chart forms a handy means of converting from impedances to admittances. Once you know an impedance, you can find the corresponding admittance by going halfway around "r=constant" circle to the diametrically opposite point and reading the values at that point as conductance and susceptance. This operation amounts to traversing a quarter wavelength the section of line whose characteristic impedance is 1 and finding an impedance which is just the reciprocal of the impedance from which you started, that is, equal to its admittance, numerically. Points B and C in figure 213 are examples of a pair of diametrically opposite points.

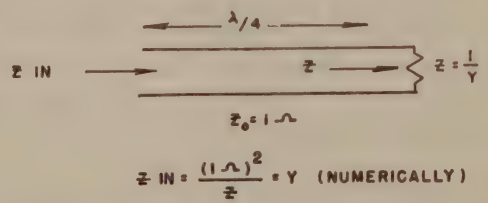


FIGURE 214.—Quarter wavelength transformer.

PRECAUTIONS.—In transforming impedances and susceptances down a line one assumes that the line does not introduce any reflections or phase distortions. If any section of the line does not satisfy these conditions, a transformation through this section will not be valid, unless the reflection or phase distortion is taken into account. Phase distortions occur at sharp right-angle bends in coaxial lines and wave guides and at resonant structures

of appreciable length, such as TR boxes. A very thin resonant structure, such as a resonant slot in a thin sheet of metal across a wave guide, will not introduce any phase distortion, and one can measure wavelengths down the line past it as though it were not there. Choke-flange joints and connectors for coaxial lines introduce no measurable phase distortion. Broad-band stub supports introduce a slight phase distortion but it is so small (about 1 mm. maximum) that it can usually be neglected. The electrical distance around curved lengths of coaxial or wave-guide lines can be measured if the mean lengths are used. A dodge to measure the electrical lengths around *reflectionless*, sharp-right-angle corners is shown in figure 215.

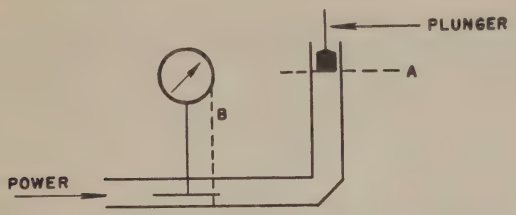


FIGURE 215.—Set-up for measuring the electrical length around a reflectionless corner.

Suppose we wished to transform an impedance to point A. Place a plunger at A. An infinite standing wave ratio will be set up with a minimum at the face of the plunger. Therefore any minimum point B along the slotted section will be an integral number of half wavelengths from A. After the point B has been found, all measurements can be made with respect to it, because any impedance at A will transform to itself at B.

Another method often used for locating an inductive diaphragm in a wave guide for matching is the use of the sliding diaphragm. This consists of a short piece of thin-walled wave guide which just fits inside a standard wave guide. An inductive diaphragm of the appropriate size (as described in Section IV, 3a) is made a part of the smaller wave guide. The device is inserted in the line at the point where the match is to be made and then slid along until the standing wave is a minimum. It can then be removed and a permanent diaphragm soldered in place at the position thus determined.

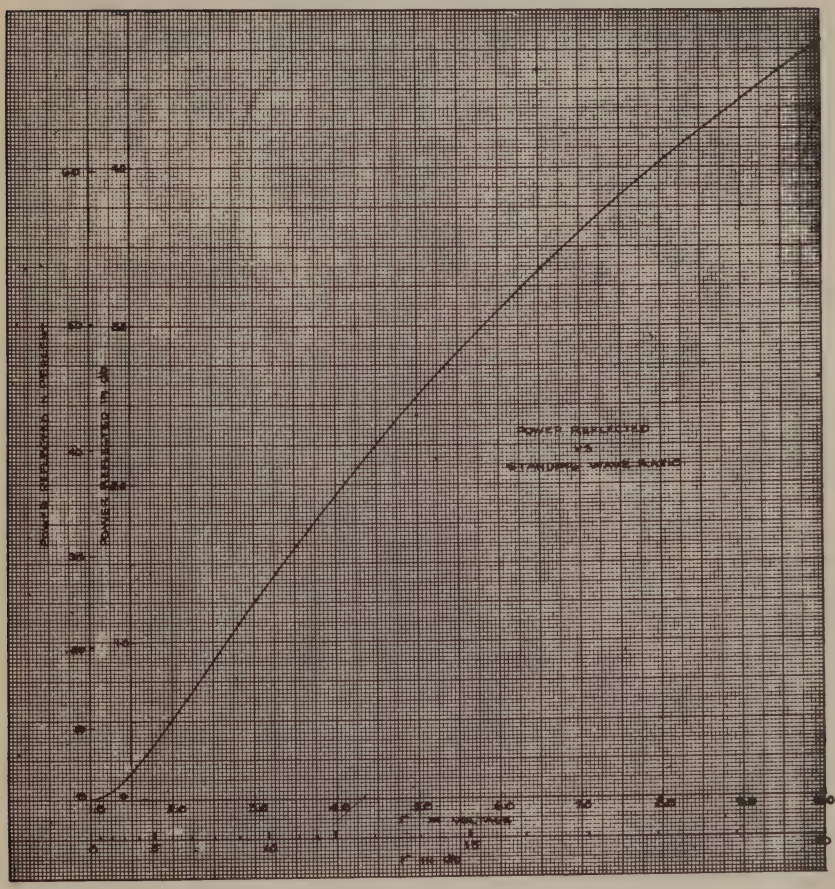
It is to be noted that it is not necessary for the sliding diaphragm to have exactly the right susceptance in order to determine the position accurately. Thus, a set of some half dozen such diaphragms could be built which would cover any standing wave ratio from one to two in voltage.

There is still a third method for accurately measuring electrical length when going around a corner or along a piece of wave guide that is too long to allow a precise calculation of phase shift. An advantage of this method is that it may be used when it is not possible to disconnect the far end to use the plunger or the sliding diaphragm.

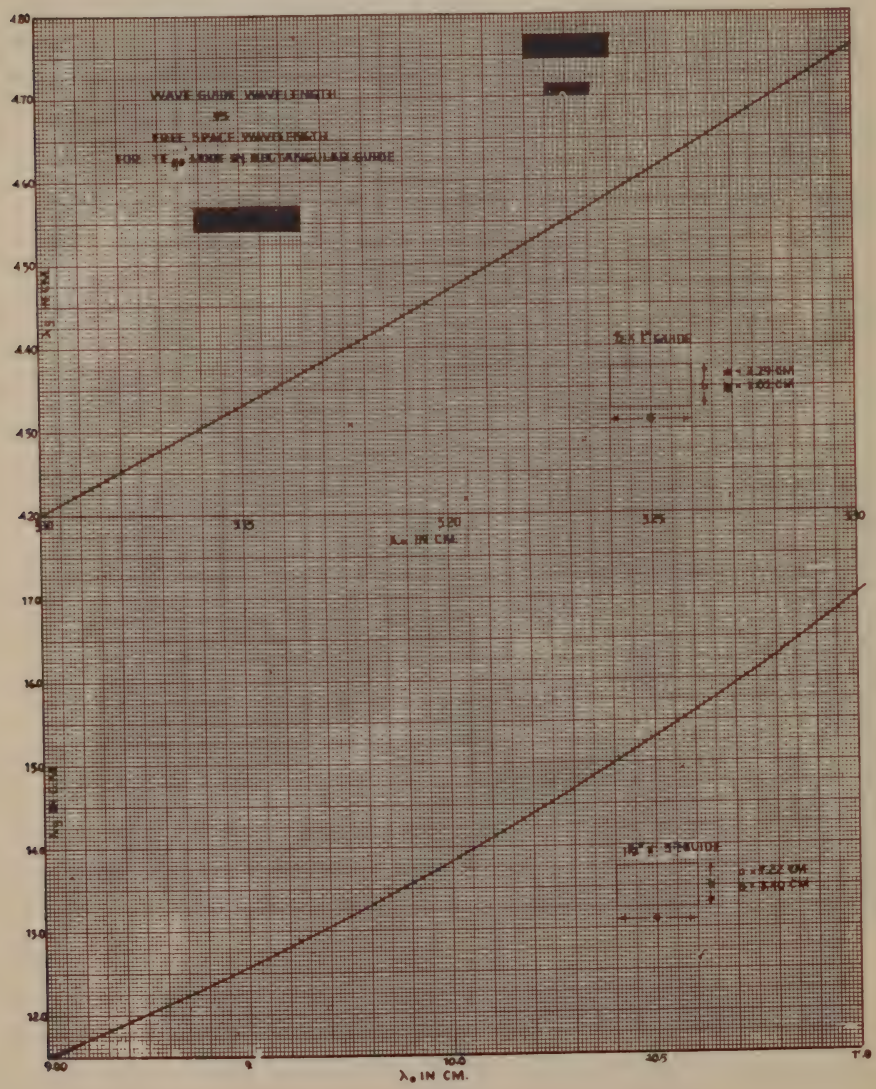
A hole is tapped for a small capacitive screw near the point where the matching is to be done and the standing wave pattern is measured for two or three insertions of the screw. Assuming some reasonable value of length of line to the position of the screw, we can plot the corresponding admittances at the assumed position of the screw on a P. H. Smith chart. If the correct line length was chosen, the points should all lie on the same circle of constant conductance. If not, the points (plotted on tracing paper) can be rotated about the origin until they do fall on such a circle. The amount and direction of rotation give the correction to the assumed line length. Then the matching procedure can be carried out using the position of the screw as the known reference point.

V. APPENDIX

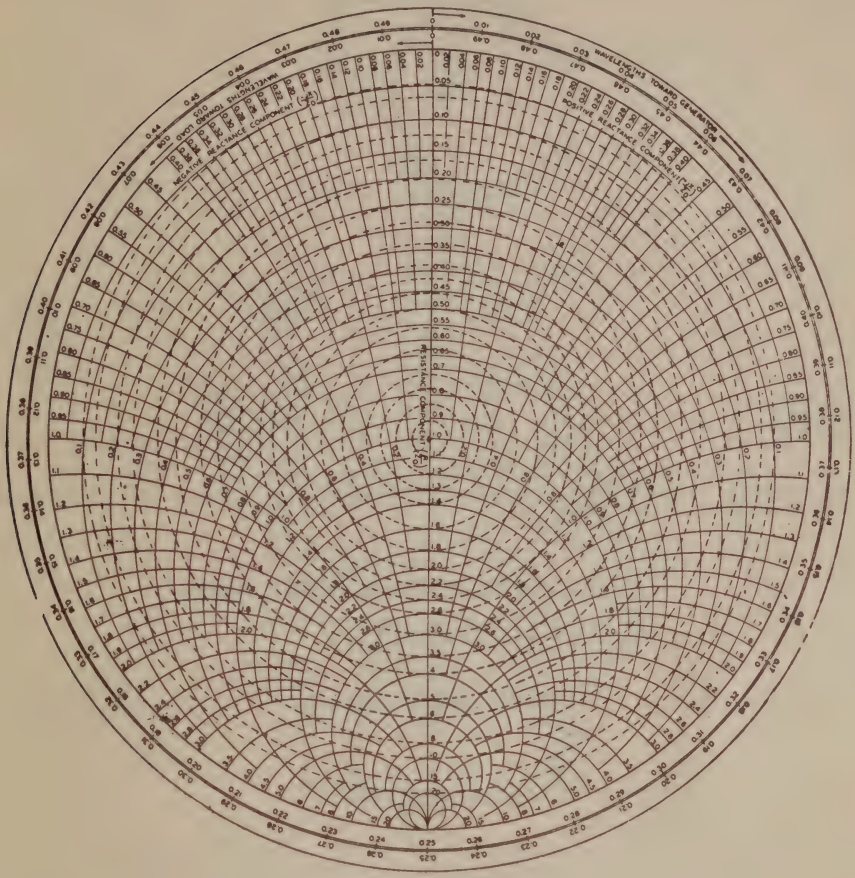
1. POWER LOST IN REFLECTED WAVE



2. GUIDE WAVELENGTH VS. FREE SPACE WAVELENGTH



3. CIRCULAR IMPEDANCE CHART



4. COAXIAL LINE DATA

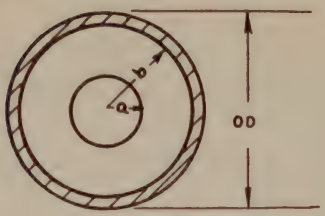


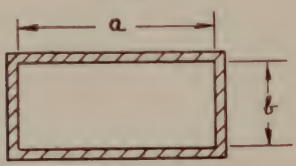
TABLE I

O. D.	Wall	$2b$	$2a$	$Z_0=138 \log_{10} b/a$	Support	Attenuation*
	Inch	*Inch*	*Inch*	*Ohms*		*db/m*
5/16″=0.3125″	0.025	0.2625	0.125	44.4	Bead	0.47
1/2″=0.500″	.032	.4375	.1875	50.6	Stub	.49
5/8″=0.625″	.035	.555	.250	47.8	Bead	.22
7/8″=0.875″	.032	.811	.375	46.4	Stub	.15
1 3/8″=1.625″	.049	1.527	.625	53.4	Stub	.08

*Calculated for $\lambda=10$ cm., except for the ½-inch line, where $\lambda=3.2$ cm. was used. Losses in beads and stubs are neglected. Material is brass.

5. RECTANGULAR WAVE GUIDE DATA

TABLE II



Size (O. D.)	a (cm.)	b (cm.)	λ (cm.)	$TE_{1,0}$ Mode			$TE_{2,0}$ Mode
				λ_g (cm.)	λ_c (cm.)	Atten.* (db/m)	λ_c (cm.)
½″ x 1″ (0.050″ wall)	2.29	1.02	3.20	4.48	4.57	0.23	2.29
			3.30	4.77			
			3.40	5.09			
5/8″ x 1¼″, (1/16″ wall)	2.86	1.27	3.20	3.86	5.72	.14	2.86
			3.30	4.04			
			3.40	4.23			
1½″ x 3″ (0.080″ wall)	7.22	3.40	9.1	11.70	14.44	.04	7.22
			9.8	13.33			
			10.7	15.92			
1 5/32″ x 2 27/32″ (3/64″ wall) (Toll Ticket)	6.98	0.952	9.1	12.05	13.96	.24	6.98
			9.8	13.8			
			10.7	16.7			

*Attenuation calculated for brass wave guides.

6. ROUND WAVE GUIDE DATA

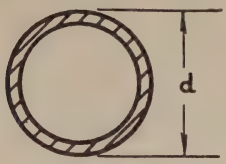


TABLE III

Size (I. D.)	d (cm)	λ (cm)	$TE_{1,1}$ Mode			$TM_{0,1}$ Mode			$TE_{2,1}$ Mode	
			λ_g (cm)	λ_c (cm)	Atten. (db/m)	λ_g (cm)	λ_c (cm)	Atten.* (db/m)	λ_g (cm)	λ_c (cm)
1 5/16"	2.38	3.20	5.17		0.18					
		3.30	5.54	4.06			3.11			2.45
		3.40	6.15							
0.930" (Al.)	2.36	3.20	5.27		0.12					
		3.30	5.75	4.03			3.08			2.43
		3.40	6.36							
1 3/16"	3.02	3.20	4.08		0.08	5.49		0.135		
		3.30	4.30	5.15		6.04	3.94			3.11
		3.40	4.52			6.73				
2 1/2"	6.35	9.1	16.8							
		9.8	23.0	10.83	0.12		8.29			6.54
		10.7	70.0							
2 3/4"	6.99	9.1	14.1			11.80				
		9.8	17.2	11.92	0.047		9.12			7.19
		10.7	24.2							
3"	7.62	9.1	12.8			22.4				
		9.8	14.9	13.00	0.030	56.4	9.95			7.85
		10.7	18.8							
4"	10.16	9.1	10.7			12.5			18.5	
		9.8	11.9	17.33	0.012	14.5	13.27	0.02	28.0	10.46
		10.7	13.6			18.1				

*Attenuation calculated for brass wave guides (except for the case of the 0.930 inch aluminum guide).

INDEX

	Page
admittance measurements	176
amplifiers	106
"K-1"	107
Type "O"	109
Type "P"	109
antenna matching	83
anti-TR boxes:	
10 cm	54
3 cm	63
attenuation:	
coaxial line	184
flexible cables	12
measurements	173
wave guide	24, 184, 185
attenuators	153
cut-off	153
dissipative	155, 157
balanced bridge	145
bead support	2, 4
bends:	
coaxial line	2
wave guide	29
"black box"	150
bolometer	102
bolometer bridge	145
break-down:	
coaxial line	4, 7, 23
flexible line	15
wave guide	23
broad-banded stub support	3
"bullet"	6
cables	12
cable loss	173
carbonized rods	156
characteristic impedance:	
coaxial line	184
flexible line	14
wave guide	25
choke-flange joint	30
coaxial lines	1
coaxial line to wave guide	44
connectors:	
stub-supported lines	5
flexible lines	16
wave guide	30
crystals	76, 97
burn-out	7, 9
conversion loss	79
equivalent circuit	78
crystal protection	55
current flow	19, 95
cut-off wavelength	18, 38
diaphragm, matching	26
dielectric in guide	38
dirt in guide	37
dry load	148
duplexing assembly	64
echo boxes	128
tuned	129
untuned	130
flexible lines	12
flexible guide	29
F-m bench	65, 170
frequency standards	123
"gozanywhere" stub support	4
guide wavelength	19, 184, 185
impedance measurements	176
Johnson meter	142
Johnson noise	78
Klystron	68, 131, 132
leakage power	56
Lighthouse tube	69
loads	148
coaxial	148
wave guide	150
loss measurements	173
magnetron coupling	48
matching devices	112
matching diaphragm	26
McNally tube	69
mixers	71
crystal	72
fixed tuned	73
pencil	72
pot	72
Pound	74
Sharpless	74
tube	71
modes:	
coaxial	1, 20
wave guide	18
rectangular	19
round	22
modulation	133
noise	71
noise figure	71, 78
obstacles in guide	26
oscillators	67, 131
local	67
test bench	131
platinized rods	156
plug attenuator	159
plunger, wave guide	32
power divider	120
power measurers	135
high level	135
low level	145
power supplies	134
pressurizing	86

	Page
probes	96
crystal	97, 101
bolometer	102
thermocouple	106
pulling	49
Q:	
measurements	66, 167, 172
of TR boxes	66
quarter wavelength transformer	8, 178
resistive (resistor) strip	151, 157
resonant sections	25
resonant structures	28
R-f envelope indicator	162
R-f head	84
"radome"	88
ringing time	128
rotating joint:	
coaxial line	9
contact type	9
capacity type	9
wave guide	47
Salisbury screen	150
Samuel tube	106, 133
sand load	148, 150
Shepherd tube	69
sizes of lines	7, 184
skin depth	38
slotted sections	91
coaxial	92
wave guide	94
space matcher	151
spectrum analyzer	159
spinner housing	88
standing waves	91, 181
stub support	1
broad-banded	3
"gozanywhere"	4
taper:	
coaxial line	8
wave guide	40
terminations	148
test benches	158, 159
test sets	165
$TE_{1,0}$-$TM_{0,1}$ transformer	42
thermistor bridge	146
T joints	32
frequency sensitivity	37
transformer:	
coaxial-to-wave guide	44
double quarter wavelength	8
matching	112, 116
quarter wavelength	8
rotary	42
tapers	39
$TE_{1,0}$-$TM_{0,1}$	42
wave guide	39
TR boxes	49
equivalent circuits	52
purpose and theory	49
10 cm	53
3 cm	58
tuners	112
screw	117
slug	113, 117
stub	112, 116
three-screw	118
turn-on	56
unbalanced bridge	146
variable impedance transformers	112
water load	135
coaxial	135
wave guide	142
wave guides	17, 26
wave meters	121
Q of	128
use	126
windows	26
capacitive	26
inductive	26
wood load	151

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